FREQUENCY MODULATED RADAR

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PREFACE

A program of research and development in the field of frequency-modulated radar was initiated by RCA Laboratories in 1938 and, beginning in 1941, the program was greatly expanded, with support provided by the United States Navy through a series of contracts. This program is believed to represent a considerable proportion of the total work that has so far been done on f-m radar.

The present book was originally prepared as a final report to the Navy under contract NXsa-35042. Concepts and practices in the field of frequency-modulated radar differ considerably from those in the better known field of pulse radar, so it is our hope that publication of this report as a book may usefully fill a real gap in the fast-growing literature of radar.

The material here presented is intended to be complete enough to be of value to readers entirely unfamiliar with the specialized subject of f-m radar. In order to hold the presentation down to a reasonable length, however, it has been assumed throughout that the reader is thoroughly familiar with the normal techniques of radio engineering, and has some familiarity with pulse radar and with servo mechanisms. Special principles and circuits typical of f-m radar are discussed in detail even though, like cycle-rate counters and wave-shaping circuits, they are also used in other fields. Some phases of the work have been described more fully in special reports submitted to the Navy from time to time; references to these are given in the text.

An attempt has been made to use consistent, unambigous and readily recognizable notation throughout. Convenience has, however, dictated double use of some symbols where the context is such as to avoid confusion. For example, T represents a time interval in discussion of bomb kinematics or a temperature in discussion of noise or rocket firing. The notation used is listed at the end of each chapter; double use of any symbol is indicated in these lists.

Production equipment described herein was all based on engineering prototypes developed by RCA Laboratories.

Design of the **AN/APN-1* and *AN/APG-4* systems for production was done by engineers of the RCA Victor Division of the Radio Corporation of America, under separate contract. Production design of the *AN/APG-17* equipment was done by the Admiral Corporation, which also worked on a production design of *AN/APG-6*. The stabilizing equipment used with *AN/APG-6* was developed by the C. L. Norden Co., Inc. Calibrating devices were developed and produced both by the RCA Victor Division and by the Raytheon Manufacturing Company.

Many persons within the Bureau of Aeronautics and the Bureau of Ships of the U. S. Navy contributed to the setting up of operational requirements which guided the entire development program, and assisted in the implementation of that program. Extensive flight testing of each equipment, both during and after development, was made possible by the cooperation of a number of naval activities. The bulk of such field work was handled through U. S. Naval Air Material Center, Philadelphia, Pa., Naval Aircraft Modification Unit, Johnsville, Pa., and Naval Air Experiment Station, Patuxent River, Md. Several other activities also took part in various special tests.

The work of a rather large number of the members of the staff of RCA Laboratories, under the general guidance of Dr. Irving Wolff, has provided the information presented. It is quite impracticable to apportion fairly the credit for individual contributions, and no attempt will be made to do so. Mention must be made, however, of the importance of the ingenuity and initiative of Mr. Royden C. Sanders, Jr., to the formative stages of the f-m radar program.

The writer wishes to acknowledge his especial indebtedness to Dr. Irving Wolff and Mr. C. C. Martinelli, both of whom have kindly reviewed the entire manuscript of this report, for much stimulating discussion and many helpful suggestions. Mr. Martinelli prepared material here presented on the *AN/APG-6(XN)*, *AN/APG-17*, and *AN/SPN-2(XN)* equipments.

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RCA LABORATORIES

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PART ONE

PRINCIPLES AND TECHNIQUES

CHAPTER I.

INTRODUCTION

1. HISTORICAL SKETCH

Echo sounding with electromagnetic waves, now known as radar, was first used at the end of 1924 for the purpose of proving the existence of a conducting layer of ionized gas in the upper atmosphere.[1*] Success in echo sounding requires that the returning echo be distinguishable from the signal originally sent out. Breit and Tuve[2] in the United States sent out a series of short bursts or pulses of radio signal, receiving delayed echos in the silent intervals between transmitted pulses. This was pulse radar. Appleton and Barnett[3] in England varied the frequency of their transmitted signal, so that the delayed echo received at any instant had a frequency different from that of the signal being transmitted at the same instant. This was frequency-modulated radar. Either method allows one to measure time lag between transmission of a signal and return of its echo and so, the velocity of signal propagation being known, to determine the distance to the reflecting object that causes the echo.

Frequency-modulated radar has been reinvented several times, usually with a view to its application for the measurement of altitude of aircraft.[4,5,6] Its importance became apparent when an altimeter of this sort[7] came into very wide use in military aircraft. Frequency-modulated altimeters were in fact used by all major combatants in World War II. The total effort so far applied to f-m radar development has however been very small, as compared to the tremendous concentration of technical effort which resulted in spectacularly rapid development of pulse radar for many military uses.

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*A list of numbered references is given in the final section of each chapter.

2. SPECIAL CHARACTERISTICS

Frequency-modulated radar both transmits and receives signals simultaneously and in most cases continuously. Since this prevents the time-division antenna duplexing used in pulse radar, separate transmitting and receiving antennas have been customary. This separation of transmission and reception makes feasible the measurement of ranges as short as a few feet.

Useful output of the radio equipment is in the form of a beat note between transmitted and received signals, the frequency of the beat depending on range and speed of the reflecting object or target. When several targets are present, the resulting complex beat must be examined with a wave analyzer in order to observe their respective ranges. Because of the finite transient-response time of selective circuits, this is inherently a somewhat slow process.

Target speed has a strong effect on the beat frequency produced. This makes frequency-modulated radar especially suitable for applications requiring measurement of both distance and speed. It also offers advantage in cases where moving targets must be distinguished from stationary ones.

Where a single reflecting object is predominantly responsible for the echo signal received, the beat-note output may be utilized by remarkably simple means to perform automatic control functions, in accordance with target distance and speed together or either quantity alone. Altitude of aircraft in level flight has been successfully controlled in this way, as has release of missiles.

Signal-to-noise ratio and therefore maximum range is of the same order for frequency-modulated and pulse radar systems of equal average power and time of response, but the f-m system has the advantage of not being required to handle large peak power. Range resolution and ability to determine small increments is roughly the same in both systems for the same radio-frequency bandwidth. Because it has facilitated the practical use of very wide frequency bands, f-m radar has had the advantage in range resolution also.

3. USES

Radar of the frequency-modulated type has so far been used predominantly in cases in which only one important reflecting object is illuminated by the transmitted signal. The major use has been for aircraft altimeters, with the reflection occurring at the surface of the ground.

Bombing from low altitudes has been found to present a type of kinematical problem to which f-m radar is well adapted to give a fully automatic solution. Isolated vessels on the surface of open water represent single targets against which very accurate results have been secured in this way. Most special systems so far proposed or built have been intended to solve various aspects of the problem of attack against surface vessels by aircraft. Use against aircraft has also been proposed. Operation has been mainly at radio frequencies near 450, 1500 and 4000 megacycles per second.

One search system using a spectrum analyzer has been tested against multiple targets. The possibilities of f-m search radar have also been studied to some extent in England.

4. REFERENCES

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7. *AN/ARN-1*, **AN/APN-1*. See Chapter VI.

CHAPTER II.

PRINCIPLES OF OPERATION

1. Determination of Distance

a. *Distance as a Time Delay.* The relation between target distance and time delay is basic for all radar ranging. If a disturbance propagated at a constant speed c travels to a target at distance or range R and an echo returns, the echo will be received a time τ after transmission of the signal, where

$$\tau = 2R/c. \tag{II.1}$$

The speed of electromagnetic-wave propagation in normal sea-level air is 299.69 megameters per second or 983.24 feet per microsecond, and this value should be used for c in any numerical calculations.

b. *Effect of Time Delay on a Signal of Variable Frequency.* Let a radio signal be transmitted at a frequency F increasing uniformly with time, as indicated by the full-line graph of Fig. II.-1. The slope of this line is the rate of change of frequency with time dF/dt or, for brevity, the *frequency rate* $\dot{F}$. If a portion of this signal is reflected by a stationary target and returns to the transmission point after a time delay τ, the received frequency will vary as shown by the dashed-line graph of Fig. II-1.

A signal transmitted at any time t_1 with frequency F_1 will be received after reflection at time $t_1+\tau$, still with frequency F_1. But the signal being transmitted at time $t_1+\tau$ will have a new frequency F_1' where, as is evident from the figure,

$$F_1' = F_1 + \dot{F}\tau. \tag{II.2}$$

A frequency difference $\dot{F}\tau$ therefore appears between signal transmitted and signal simultaneously received after reflection from a stationary target.

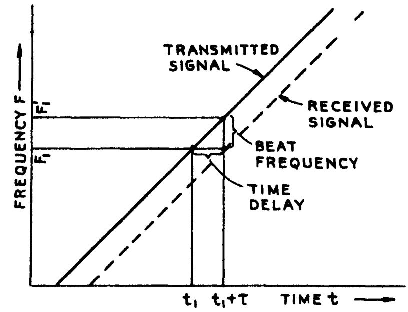

Fig. II.-1. Effect of time delay on signal with uniformly varying frequency.

c. *Range Frequency f_R*. The frequency change from F_1 to F_1' does not depend on t_1 or F_1. Therefore a beat note formed by the receiving equipment between the transmitted signal of uniformly varying frequency and the returned echo with fixed delay will have a constant frequency. Since this beat frequency depends, through time delay τ, on the range R of the reflecting target, it will hereafter be referred to as *range frequency* f_R and, from equations (II.1) and (II.2) is given by

$$f_R = (2/c)\dot{F}R. \tag{II.3}$$

2. Determination of Speed

a. *Doppler Effect of Moving Transmitter*. Electromagnetic waves emitted from an antenna in empty space move outward in all directions from that antenna with speed c, independently of any motion of the antenna. If a stationary antenna oscillates electrically at a frequency F or period $1/F$, the disturbance radiated at the peak of one oscillation will move outward a distance c/F by the time the peak of the next oscillation is radiated. Fig. II.-2(a) illustrates this well known action, which gives rise to radiated waves of length

$$\lambda = c/F. \tag{II.4}$$

Now consider an antenna moving with speed S. Waves emitted in the direction of the motion will move away a distance c/F during one oscillation period, but in the same time the antenna will move a distance S/F in the same

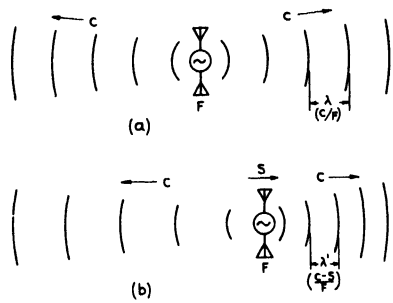

Fig. II.-2. Doppler effect of moving transmitter.

direction. This tendency of the antenna to catch up on the emitted radiation results in compression of the emitted waves to a net length

$$\lambda' = (c - S)/F, \tag{II.5}$$

as illustrated in Fig. II.-2(b). Radiation emitted in the opposite direction of course has its waves stretched rather than compressed. Dependence of radiated wave length on motion of the wave source is called *Doppler effect*, in honor of its discoverer.

b. *Doppler Effect of Moving Receiver.* An electromagnetic disturbance of wave length λ passing a stationary receiving antenna at speed c induces in the antenna an electromotive force which completes one cycle of variation in the time λ/c required for one wave to pass. The frequency of the induced e.m.f. is therefore

$$F = c/\lambda. \tag{II.6}$$

If the receiving antenna is moving directly into the arriving disturbance with a speed S, each wave will pass in the decreased time $\lambda/(c+S)$ and the induced e.m.f. will therefore have its frequency increased to

$$F' = (c + S)/\lambda. \tag{II.7}$$

Conversely, the frequency induced in an antenna moving away from the source of waves is decreased.

Dependence of apparent frequency of a wave on the speed of the observer is also a *Doppler effect*. Standard homely examples of the two effects, due respectively to motion of source and observer, are the change in pitch of a loco-

motive whistle heard by a bystander at a railroad crossing as the locomotive passes him, and the change in pitch of the crossing bell heard by a train passenger as he passes the crossing.

c. ***Doppler Effect for Radar.*** A radar transmitter carried by an airplane flying at speed S toward a stationary target and operating at frequency F sends toward the target waves of reduced length λ' [see equation (II.5)], and the target reflects these waves back toward the airplane without change of wave length. The airplane's radar receiving antenna, moving into the reflected signal of wave length λ', has induced in it an e.m.f. with frequency increased according to equation (II.7).

Combining equations (II.5) and (II.7), the received frequency is

$$F'' = F(c+S)/(c-S). \tag{II.8}$$

This is equivalent to

$$F'' = \left\{ 1 + \frac{2S}{c} + \frac{2S^2}{c(c-S)} \right\} F, \tag{II.8a}$$

as may be seen by clearing fractions in (II.8a). Since the fastest projectile now used, the V-2 rocket, travels at a speed which is only 1/200,000 of c, a practically perfect approximation to the received frequency F'' is always given by

$$F'' = F + (2/c)SF. \tag{II.9}$$

If the transmitter and receiver are stationary but the reflecting target moves toward them at speed S, the target has induced in it an electromotive force of increased frequency and re-radiates toward the receiver at reduced wave length. But successive use of equations (II.4,7,5 and 6) again gives exactly the result (II.8). In other words, it makes no difference whether radar, target, or both are moving: only their relative speed along the line joining them counts.*

*The above reasoning is oversimplified, as relativity theory is required to give a really adequate explanation of Doppler effect on electromagnetic waves. However, the final result of equation (II.8) is exactly correct and the simplified physical picture is very straight-forward and easily grasped.

Fig. II.-3 shows the effect of relative target motion on a reflected signal with sweeping frequency. As in Fig. II.-1, the full and dashed lines represent respectively transmitted signal and signal received from a stationary target. The dotted-line graph represents the variation of frequency with time of a signal received after reflection from a moving target (the effect of target motion is shown greatly exaggerated). A signal transmitted at time t_1 with a frequency F_1 returns at time $t_1+\tau$ with a frequency F_1'', while

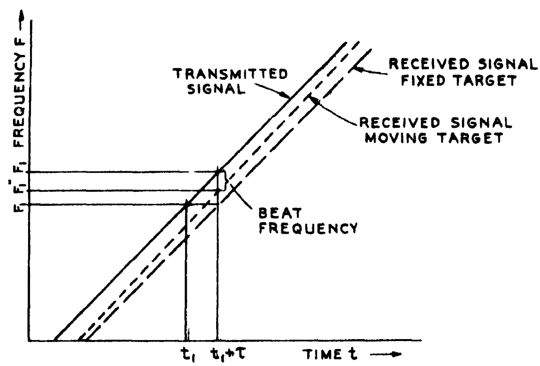

Fig. II.-3. Effect of target motion.

the frequency of transmission at time $t_1+\tau$ is F_1' and the difference between frequencies simultaneously transmitted and received is $F_1' - F_1''$.

d. *Speed Frequency* f_S. From (II.1), (II.2) and (II.9), the beat-note frequency f resulting from combination at time $t_1+\tau$ of transmitted signal and signal reflected from a moving target, having frequencies F_1' and F_1'' respectively, is

$$f = (2/c)\dot{F}R - (2/c)F_1 S. \qquad \text{(II.10)}$$

The first term of this expression is just the range frequency f_R discussed earlier. The second term depends on relative target speed only and not on range; it will hereafter be referred to as *speed frequency* f_s, so that

$$f_S = (2/c)FS. \qquad \text{(II.11)}$$

This is, of course, simply the Doppler frequency shift.

3. ALTERNATIVE PICTURES

a. *Quasi-Steady State.* Frequency rates used in most f-m radar are such that the emitted frequency changes at most a small fraction (0.01) of one per cent in the time taken for a signal to travel from transmitter to target and back to receiver. Under these conditions it is permissible to consider that the fields between radar and target behave at each instant essentially as if the transmitter had been radiating the same frequency for a considerable time, and to say that a quasi-steady state exists. This permits us to imagine the electromagnetic waves emanating directly from the radar and those radiating from a reflecting object as combining at any instant to give a stationary pattern of wave amplitude in the region around the reflecting object. Such a condition leads to an alternative concept of f-m radar operation that is particularly convenient for study of certain special properties.

It is the existence of a quasi-steady condition that permitted use of the simple viewpoint of Fig. II.-1. So long as the change in instantaneous frequency during each radio-frequency cycle is sufficiently slight, it is permissible to regard a frequency-modulated signal as having a single but slowly varying frequency. In the case of more drastic modulation, this picture must be used only with extreme care; it may then be more appropriate to regard the signal instead as a group of constant-frequency Fourier components. It is fortunate for ease of understanding that f-m radar operation is usually of the quasi-steady type.

b. *Standing-Wave Pattern.* Along the line joining transmitter and reflector, waves approaching the reflector will be so related in phase to waves leaving it as to give a maximum resultant amplitude once in each half wave length of distance from the reflector. Similarly, there will be one minimum of resultant amplitude per half wave length, as indicated in Fig. II.-4. In range R there will be $R/(\frac{1}{2}\lambda)$ complete standing-wave amplitude cycles. From the relation (II.4) between wave length λ, frequency F and propagation velocity c, the number N of standing-wave

cycles in range R is seen to be

$$N = (2/c)RF. \qquad \text{(II.12)}$$

c. ***Vector Representation.*** Still another picture may be obtained if signal voltages in the radar receiver are represented by rotating vectors, as is often done in describing behavior of electrical circuits. Let the signal reaching the receiver directly from the transmitter be taken as a reference. This signal will then be represented, as in Fig. II.-4, by a vector e_1, always of unit length and always directed horizontally to the right. The signal returned by the target will be represented by a vector e_2, of length and direction corresponding to echo-signal amplitude and phase relative to e_1. Resultant signal in the receiver is represented by a vector e_3, the sum of e_1 and e_2. All vector configurations are considered to be observed from a reference frame which rotates clockwise at the frequency of the reference signal.

Suppose the direct and target-reflected signals at the receiver to be in exactly opposite phase for zero target range, as they would be at the surface of a perfectly reflecting target. Let both signals reach the receiver from the transmitter, for zero range, by paths of exactly equal length. Then e_2 will lag e_1 in phase at any particular instant, for any target range R, by an angle ψ. Phase lag ψ will be just π radians or 180 degrees more than the angle through which the observer's reference frame has turned during a time interval $2R/c$ immediately preceding the instant in question. For a quasi-steady condition, ψ will be $2\pi(N+\frac{1}{2})$ radians or $360(N+\frac{1}{2})$ degrees, where N is the number of standing waves given by equation (II.12).

d. ***Effect of Changes.*** As the transmitted frequency is slowly increased, more and more standing-wave amplitude cycles must appear within a fixed range R between radar and target. For each full wave added to the standing pattern, a fixed receiver near the transmitter will find one cycle of amplitude modulation of the total received signal. This behavior may be thought of as a deformation of the standing-wave pattern, with new waves entering the pattern at the transmitter as indicated in Fig. II.-4. The corresponding relative rotation of the reflected-signal

vector e_2 is also shown in the figure. Thus, as the number N of standing waves alters, the total received signal will undergo an amplitude modulation of frequency dN/dt. From equation (II.12), dN/dt for constant range is seen to be

$$(\dot{N})_R = (2/c)R\dot{F} = f_R. \tag{II.13}$$

This is exactly the range frequency f_R (II.3) found in a different way in section 1c above. Rate of change with frequency of relative phase ψ is evidently proportional to range.

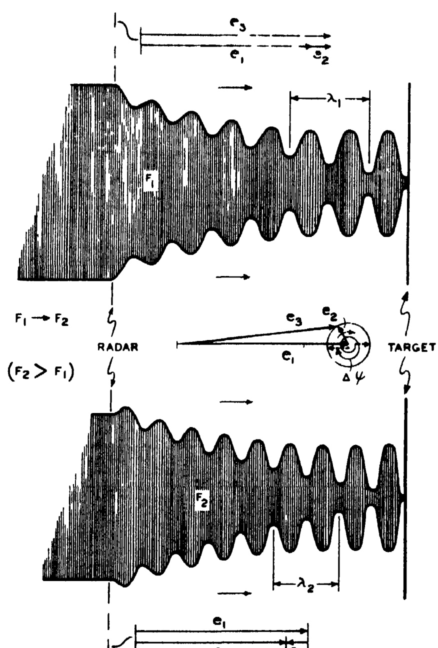

Fig. II.-4. Standing-wave patterns and vector diagrams for signal of varying frequency.

For a fixed radio frequency, the standing-wave pattern is fixed in space with respect to the reflecting object or target. Therefore, if transmitter and receiver move toward the target, one cycle of amplitude modulation of the resultant received signal (one full rotation of vector e_2 relative to e_1) occurs for each wave of the standing pattern traversed— that is, for each half wave length moved. The frequency of the received amplitude modulation dN/dt is,

for constant radio frequency,

$$(\dot{N})_F = (2/c)\dot{R}F. \qquad \text{(II.14)}$$

This is of exactly the same magnitude as the speed frequency f_S (II.11) found in section 2d above (since range decreases during approach, closing speed S is just the negative of rate of change of range $\dot{R}$). Rate of change of relative phase ψ with range is evidently proportional to frequency.

If both range and frequency vary, differentiation of equation (II.12) shows that the total frequency of the received amplitude modulation, due both to expansion or contraction of the standing-wave pattern and to motion of the radar through the pattern, is

$$\dot{N} = (2/c)\dot{F}R + (2/c)F\dot{R} = f_R - f_S. \qquad \text{(II.15)}$$

Rate of change of signal phase, $\dot{\psi}$, is of course just $2\pi\dot{N}$ radians per second.

e. *Equivalent Viewpoints*. Behavior of a standing-wave pattern thus provides for quasi-steady conditions an alternative to the time-lag and Doppler-effect explanation of the beat note produced by f-m radar. Still another explanation may be given in terms of electrical vectors. Each physical picture is quite correct and the choice among the three methods of explanation is simply a matter of suitability to a particular problem or of preference, since they are substantially equivalent.

One property of the combined direct and reflected radar signals is hidden in the simple standing-wave envelope picture. The resultant signal actually varies in phase as well as in amplitude when either the radar moves with respect to the target or the transmitted frequency varies. This phase variation rather than the amplitude variation might in principle be used to measure range and speed.

For practical reasons, it has proved desirable in actual f-m radar systems to minimize space transmission directly to the receiving antenna. Instead of such transmission, a controllable coupling from transmitter to receiver is provided within the equipment. The essential features

of the variation of receiver output are, nevertheless, still given correctly by the simple standing-wave picture, which may therefore be retained wherever convenient. The vector picture is more complete, since it can allow for any kind of direct-signal coupling and displays phase as well as amplitude variations of the resultant signal.

4. PERIODIC FREQUENCY MODULATION

a. *Need for Periodicity.* Only the simplest type of frequency modulation, a continuous uniform variation of frequency, has been used above as the basis for deriving the characteristic properties [eq. (II.10) or (II.15)] of f-m radar. The results show that with such modulation it is not convenient to determine separately range and closing speed from the beat-note frequency of the radar output. Also, it is rather impractical to vary the transmitted frequency uniformly in one direction for any very long time. To meet these practical objections without unduly complicating beat-note properties, it is necessary to use a periodic variation of the transmitted frequency.

b. *Triangular Modulation.* A particularly simple form of periodic frequency variation has proved to be especially useful as well. In this type of modulation, the magnitude of the rate of change of frequency remains constant but the direction of the change is periodically reversed. The frequency alternately increases and decreases uniformly for equal periods of time. Fig. II.-5(a) illustrates the variation of frequency rate $\dot{F}$ with time for such modulation, while the solid-line graph of Fig. II.-5(b) represents the time variation of transmitted radio frequency F. This will be referred to hereafter as triangular or symmetrical-sawtooth frequency modulation.

Limited study of the frequency spectrum or side-band pattern of a signal modulated in this way, as determined by Fourier analysis, has not revealed any strikingly interesting properties. Therefore, no such analysis will be presented here. It seems much more productive to regard the f-m radar signal as substantially a pure or single-frequency sine wave, subject to a relatively slow alteration of frequency.

c. ***Reflection from Stationary Target.*** The broken-line graph of Fig. II.-5(b) represents the frequency of the signal received after reflection from a stationary distant object. Time variation of the beat-note frequency f is given by the difference in ordinates between the solid (transmitted frequency) and broken (received frequency) curves, which is in turn plotted directly in Fig. II.-5(c). Except for brief departures following the instants of reversal of the transmitter frequency rate $\dot{F}$, which may be called the *turn-around points* of the modulation cycle, the beat note is constantly at the range frequency f_R defined in section 1c above.

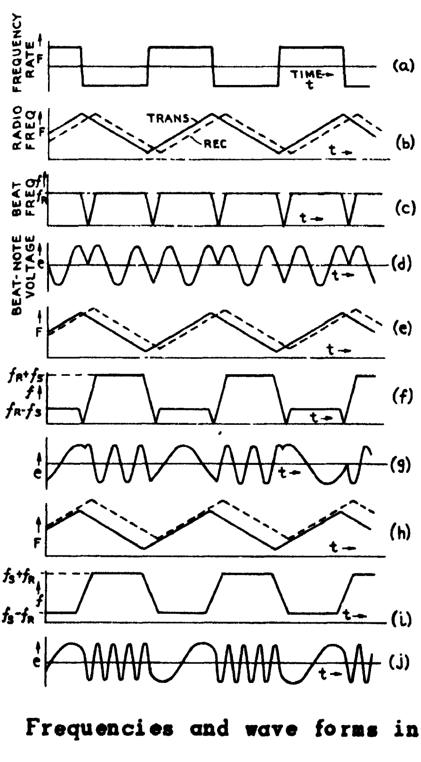

Fig. II.-5. Frequencies and wave forms in f-m radar.

d. ***Output Wave Form.*** The actual wave form typical of beat-note voltages produced by stationary targets is shown by Fig. II.-5(d). The fact that such an output wave must be produced is easily recognized by consideration of the standing-wave pattern. Rectified receiver-output voltage measures the resultant amplitude of the combined

transmitted and reflected signals, and its variations represent a partial profile of the sinusoidal standing-wave pattern formed in space by combination of these signals.

As the transmitted frequency increases, the number of standing waves between radar and target increases, so that the standing-wave pattern slides past the receiver toward the target. The receiver output varies to delineate the profile of the waves and fractions of waves of amplitude variation that pass over it. Similarly, when the transmitted frequency decreases, the standing-wave pattern reverses its motion and slides past the receiver away from the target. The receiver output again shows the form of the standing amplitude waves that pass over it.

At modulation turn around, the position of the receiver in the standing-wave pattern does not change but the motion of the pattern reverses. At each turn-around point the receiver output voltage therefore does not change but the direction of its variation reverses. This gives rise to the peculiar output wave of Fig. II.-5(d), in which the wave form during each half cycle of the frequency modulation is the mirror image of that for the adjacent half cycles. This type of output, which is substantially a sine wave of constant frequency during each half cycle of modulation but suffers a rapid and perhaps large phase change at the end of each such half cycle, is a striking characteristic of f-m radar in many conditions of use. The magnitude of the phase change depends on the particular phase of the standing wave at which the receiver happens to be located at the instant of modulation turn around. From the point of view of transmission time and beat-note formation, the phase change results from the fact, easily seen in Fig. II.-5(b), that at turn around transmitted and received frequencies change places. That is, at turn around the sign of the frequency difference changes.

Variation of resultant-signal amplitude according to Fig. II.-5(d) is also easily explained in terms of the vector diagrams of Fig. II.-4. Periodic frequency modulation of the transmitted signal results in periodic variation of phase lag ψ of the target-reflected signal component at the receiver, relative to the component arriving directly from the transmitter. This periodic component-phase variation

results in cyclical modulation of both amplitude and phase of the resultant signal, at one cycle per complete phase rotation. Component-phase lag varies linearly with both transmitted-signal frequency and target range, as indicated by equation (II.12). Total sweep of variation of ψ depends upon range and total change of transmitted frequency in modulation, while the general region of phase variation depends upon range and mean transmitted frequency.

Phase jumps of resultant-signal amplitude variation at modulation turn around are not to be confused with the smooth variation of radio-frequency echo-signal phase ψ. The slope discontinuities of Fig. II.-5(d) represent reversals of the sense of relative-phase rotation of the echo-signal vector e_2.

The range-beat signal of Fig. II.-5(d) is obviously periodic at modulation frequency f_m, so long as the target is absolutely stationary and the frequency modulation of the radar is truly periodic. Thanks to the phase jumps at turn around, this is so even when the range-beat frequency f_R is not commensurable with f_m. Fourier analysis of any stationary-target beat can therefore yield only components at exact harmonics of the modulation frequency, whatever the target range or (periodic) modulation form, so has not been found particularly useful.

e. *Reflection from Moving Target.* Fig. II.-5(e) shows, by its broken-line graph, the frequency of the signal received from a target moving at moderate speed, with the transmitted-frequency variation again indicated by the full-line graph. Time variation of the instantaneous frequency of the beat note resulting from mixing the transmitted and received signals is shown by Fig. II.-5(f), with the speed-frequency and range-frequency components indicated. The output wave form, shown as Fig. II.-5(g), differs from that for a stationary target only in the appearance of two distinct beat frequencies on alternate half cycles of the modulation. This difference usually results in turn around occurring at different phases of a beat-note cycle on each successive modulation cycle, so that the beat-note output is not in general simply periodic when the target is moving.

Reflection from a sufficiently rapidly moving target (or transmission of a sufficiently slowly varying frequency) produces the received-frequency variation shown by the broken-line curve of Fig. II.-5(h), with a beat-note frequency varying as shown in Fig. II.-5(i). The condition of Figs. II.-5(h) and (i) corresponds to a speed frequency greater than the range frequency, while that of Figs. II.-5(e) and (f) corresponds to range frequency greater than speed frequency. Either condition of operation is entirely acceptable, provided minimum and maximum ranges and speeds at which operation is required have suitable values. It is important, however, that the condition under which operation actually occurs be definitely known at all times and that the system be used in accordance with such knowledge. Fig. II.-5(j) shows the beat-note wave form resulting from high-speed operation. The output-voltage phase jumps at turn around are absent in this case, since the receiver always moves in one direction through the standing-wave pattern and only its speed relative to the wave pattern changes at modulation turn around.

f. *Upsweep and Downsweep Frequencies.* For symmetrical-sawtooth frequency modulation, frequency rate has always the same magnitude $\dot{F}$ but is positive on the upsweep of the modulation, when frequency transmitted is increasing, and negative on the downsweep, when frequency is decreasing. Reversals of sign of beat frequency, corresponding to reversals of sign of $\dot{F}$, represent merely phase jumps in the beat-note signal at the turn-around points or frequency limits of the modulation cycle. Therefore only the absolute magnitude of the beat frequency need be considered without regard to its sign.

Equation (II.10) of section 2 above gives the beat frequency correctly, no matter whether F is increasing or decreasing or whether range frequency exceeds speed frequency [case of Figs. II.-5(e), (f), and (g)]. or speed frequency exceeds range frequency [case of Figs. II-5(h), (i), and (j)]. The magnitude $|f|$ of the beat frequency is always

$$|f| = |2/c\dot{F}R - 2/cF\dot{S}| \qquad \text{(II.10a)}$$

for a transmitted radio frequency F varying at a rate $\dot{F}$.

During the upsweep of the triangular modulation, the frequency rate is $+|\dot{F}|$ and the beat has the *upsweep frequency* f_u given by

$$f_u = |(2/c)|\dot{F}|R - (2/c)FS| = |f_R - f_S|. \tag{II.16}$$

During the downsweep of the modulation cycle, the frequency rate is $-|\dot{F}|$ and the beat has the *downsweep frequency* f_d given by

$$\begin{aligned} f_d &= |-(2/c)|\dot{F}|R - (2/c)FS| \\ &= (2/c)|\dot{F}|R + (2/c)FS = f_R + f_S. \end{aligned} \tag{II.17}$$

If it is known from the conditions of operation that range frequency must exceed speed frequency, then

$$f_R = \tfrac{1}{2}(|f_d| + |f_u|). \tag{II.18}$$

and

$$f_s = \tfrac{1}{2}(|f_d| - |f_u|). \tag{II.19}$$

If it is known that speed frequency must exceed range frequency, then these expressions for f_R and f_S are simply interchanged. In general, f-m radar equipment has been designed to operate with range frequency definitely greater than speed frequency. Given such definite knowledge of type of operation, the two beat-note frequencies produced by continuous-wave radar with symmetrical-sawtooth frequency modulation determine separately both range and closing speed between the radar and any single target.

g. ***Radar Range Sensitivity***. Range beat-note frequency developed per unit range, f_R/R, may be called the *range sensitivity* k_R of a frequency-modulated radar. From equation (II.3),

$$k_R = (2/c)|\dot{F}|. \tag{II.20}$$

If the transmitted frequency is swept uniformly back and forth between limits F_1 and F_2 defining a frequency band of width W, with a modulation period t_m or modulation frequency f_m, as illustrated by Fig. II.-6, the magnitude of the rate of change of frequency is always

$$|\dot{F}| = (F_2 - F_1)/(\tfrac{1}{2} t_m) = 2Wf_m. \tag{II.21}$$

The radar range frequency with triangular frequency

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modulation is therefore

$$f_R = (4/c)Wf_m R, \tag{II.22}$$

and the radar range sensitivity

$$k_R = (4/c)Wf_m. \tag{II.22a}$$

With W in megacycles per second, f_m in cycles per second and k_R in cycles per second per foot,

$$k_R = Wf_m/245.8. \tag{II.22b}$$

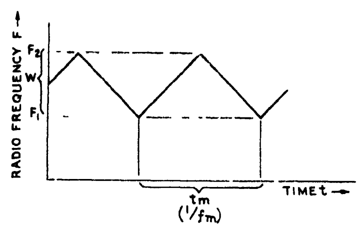

Fig. II.-6. Frequency modulation wave form.

It is an important property of this type of radar that the range sensitivity does not depend on transmitted frequency nor, separately, upon the frequency deviation or repetition frequency of the modulation. Range-beat frequency per unit of range depends only on the product Wf_m of the width of frequency band swept and frequency of sweep repetition, which has sometimes been called the *modulation product* of the transmission. Radar range sensitivity may conveniently be controlled by adjustment of either the band swept or the sweep-repetition frequency.

h. *Radar Speed Sensitivity.* Speed beat-note frequency developed per unit speed, f_S/S, may be called the *speed sensitivity* k_s of a continuous-wave radar. From equation (II.11),

$$k_s = (2/c)F. \tag{II.23}$$

Speed sensitivity varies over the frequency-modulation cycle, because radio frequency F varies. But in practice F varies only slightly, at most a few per cent, from F_0, the average over the modulation cycle of the transmitted frequency; F_0 may be considered the carrier frequency of

the transmission. Sensitivity variations are correspondingly small, so the radar speed sensitivity is for all practical purposes

$$k_s = (2/c)F_o \qquad \text{(II.23a)}$$

which is its average value.

If F is in megacycles per second and k_s in cycles per second per foot per second,

$$k_s = F_o/491.6. \qquad \text{(II.23b)}$$

One statute mile per hour is the same speed as 1.467 ft., per sec., and one nautical mile per hour or knot as 1.689 ft. per sec., so the numerical factor in the denominator of (II.23b) becomes 335.2 to give sensitivity in cycles per second per mile per hour, or 291.1 for cycles per second per knot.

It is an important property of radar speed measurement that the speed sensitivity does not depend to any significant extent on the modulation used, but only on the radio carrier frequency. Since frequency allocations and apparatus dimensions prevent any major adjustment of carrier frequency, speed sensitivity is practically a fixed characteristic of any given radar equipment.

i. *Other Types of Modulation*. Of course, many other forms of frequency modulation than the symmetrical sawtooth are possible. The latter has been discussed at length because it has been extensively developed and was used in almost all of the equipments to be described in later chapters.

Unsymmetrical sawtooth modulation may be used in special cases to make upsweep and downsweep beat frequencies equal despite the existence of target motion, by suitable control of $|\dot{F}|$, as has been done in one equipment. In the limit, with uniform frequency variation from F_1 to F_2 followed by an abrupt return to F_1, unsymmetrical sawtooth modulation does not readily permit separation of range and speed data, so is useful only against stationary targets.

When the transmitted frequency is changed from F_1 to F_2, the number of standing waves between the radar and a

stationary target will change from N_1 to N_2 and

$$\Delta N = N_2 - N_1 = (2/c)(F_2 - F_1)R. \tag{II.24}$$

This is equal to the number of cycles of variation of received output produced, so long as the frequency proceeds from F_1 to F_2 without reversals, and is seen to be independent of the exact way in which the frequency changes. If, therefore, the transmitted frequency is varied from a given value F_1 to another given value F_2 in a given time $\frac{1}{2}t_m$, the average beat frequency of the receiver output due to range only will be

$$f_R = (4/c)(F_2 - F_1)R/t_m, \tag{II.25}$$

whatever (within reason) the wave form of the modulation. These relations are, of course, modified if the target is moving.

Sinusoidal frequency modulation is simple to produce but does not lend itself readily to separation of speed and range data, at least when range frequency exceeds speed frequency. So long as only range information is required, and so long as the beat frequency can be averaged over the modulation cycle, sinusoidal modulation is permissible and equation (II.25) shows that radar range frequency is still as given by equation (II.22).

Another simple possibility is the limiting case of square-wave modulation, with frequency periodically shifting abruptly from one to the other of two fixed values F_1 and F_2. Quasi-steady concepts must be used cautiously in dealing with abrupt frequency changes, but careful consideration will show that equations (II.24) and (II.25) still apply in the case of square-wave frequency modulation. Immediately after each frequency shift, a beat signal of frequency F_2-F_1 will be present for a time interval $2R/c$, giving altogether the number of beat cycles indicated by (II.24) if the target is stationary. There will then be no further beats until the next frequency shift, after which the same sequence of events will recur.

When there is relative target motion, a speed-frequency beat will appear continuously. If the target moves toward the radar, the speed beat will increase the frequency

of the burst of range beats following each abrupt decrease of transmitted frequency and will decrease the frequency of the burst of range beats following each increase of transmitted frequency. The square-wave type of frequency modulation in its most general form seems poorly suited for separate determination of range and speed when both are present together.

If the frequency shift used is small enough so that square-wave frequency modulation changes the number of standing waves between radar and target by less than one full wave at the useful values of range, an interesting result is produced for moving targets. The received speed-beat signal will then be substantially of steady sinusoidal form, except for a fractional-cycle phase shift at each change of transmitted frequency by the square-wave modulation. The average beat frequency will measure target speed relative to the radar, and the magnitude of the beat-signal phase shifts will measure target range. For clarity of result, radio frequency, range, and speed should be so related that each beat-signal phase shift caused by modulation is completed in a small fraction of one speed-beat cycle. Also, the transmitted frequency should remain at each of its two fixed values long enough for several complete cycles of the speed-beat signal to develop, but not so long that the slight difference in beat frequency under the two conditions can confuse the result.

j. *Condition at Turn Around.* At the limits of frequency sweep in triangular modulation, where rate of change of transmitted frequency alters abruptly, the quasi-steady picture of a signal that does not change appreciably during the time $2R/c$ required for propagation must be used cautiously. A slowly varying standing-wave pattern still exists, but following modulation turn around a slight deformation propagates rapidly through the pattern. This effect may be important if propagation lag τ (which is $2R/c$) becomes a significant fraction of the duration $1/(2f_m)$ of a single modulation sweep of transmitted frequency.

Departures from quasi-steady properties as a result of time delay in signal propagation may be handled correctly if the true relation between phase and frequency of a

variable-frequency signal is borne in mind.[1] The phase ψ of an alternating signal $\sin\psi$ is always given by integration as $2\pi\int F dt$ radians, where F is the instantaneous frequency of the signal. In the usual case of constant frequency, phase angle ψ is of course found to be just $2\pi F(t-t_0)$, where t_0 allows an arbitrary starting phase.

In the vector picture of Fig. II.-4, output-signal amplitude is the magnitude of electrical vector e_3 and depends at each moment on the phase ψ of reflected-signal vector e_2 relative to transmitted-signal vector e_1. Relative phase ψ is just the difference between phase ψ_1 of the signal being transmitted at a given instant and the phase ψ_2 of the target-reflected signal being simultaneously received. Except for a fixed phase shift on reflection (π radians for an ideal reflector), ψ_2 is exactly the phase of the signal as transmitted at an instant which precedes its reception by the propagation delay τ. Relative phase ψ at any time t is therefore exactly $2\pi\int_{t-\tau}^{t} F dt + \pi$, even if F changes appreciably over the interval τ.

The results of this line of reasoning can be stated very simply in terms of time averages of variable quantities. For the standing-wave picture, number N of standing waves to be considered is at any instant exactly that given by equation (II.12), if the value of radio frequency F used in that equation is the average over a time interval τ preceding the instant in question. For the vector picture, ψ is always $2\pi(N+\frac{1}{2})$, with N determined as above. For the time-lag and frequency-shift picture, the instantaneous value of range frequency f_R is exactly as given by equation (II.3), but the value of frequency rate $\dot{F}$ used in that equation must be the average over a time interval τ immediately preceding the instant in question. Equation (II.25) for f_R as given by limits of frequency swing must be modified to allow for the fact that average transmitted frequency over an interval τ never quite reaches the extreme values F_1 and F_2 of the instantaneous transmitted frequency.

In almost all cases, the above corrections to quasi-

steady ideas are far too small to be of practical importance. Unless the existence of such corrections is realized and understood, however, paradoxical and misleading conclusions may occasionally be reached.

k. *Fixed Error.* One property characteristic of periodic modulation requires further discussion. This property is most striking when there is no target motion and the range is so short that modulation varies the number of standing waves present by less than one-half wave in all. Because the fractional change in radio frequency *F* due to modulation is usually small, this range may still be sufficient for the existence of a large total number of standing waves. Suppose the range to be constant and such that the radar receiver is at a steeply rising point of the standing-wave pattern (where direct-signal and reflected-signal vectors e_1 and e_2 are at right angles). As transmitter frequency is modulated in the way shown in Fig. II.-7(a), receiver output-voltage variation due to alternate stretching and compression of the pattern by a fraction of a wave (alternate increase and decrease of signal vector phase ψ over a range $\Delta\psi$) is of the form shown by Fig. II.-7(b). Now suppose the range to be

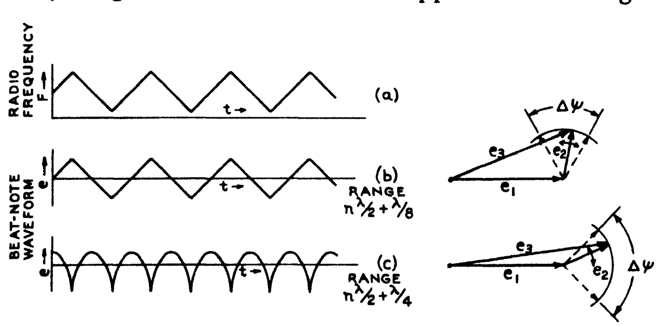

Fig. II.-7. Fixed error due to radio-frequency phase change.

changed by one-eighth wave length, so that the receiver is located at a peak-amplitude point of the wave pattern (vectors e_1 and e_2 in line). The output variation due to modulation will now have the form of Fig. II.-7(c).

The range change has been trifling, yet the output frequency has obviously changed by a factor of two! Neither of the output-signal frequencies of Fig. II.-7 is

even related to the normal f-m radar range frequency given by equation (II.3). Similar effects occur at greater ranges but are not so striking. These effects, which will be considered more fully in section 2g of Chapter IV., have sometimes been called *fixed error*. They indicate that breaking up the range-frequency beat by periodic variation of frequency rate $\dot{F}$ is not without its disadvantages. In laboratory measurements, such effects are easily observed and may prove most annoying. In actual use, however, ranges are never absolutely fixed and, particularly at the highest carrier frequencies, fixed error is averaged out by slight random fluctuations of range or by the orderly range changes due to moving targets.

1. ***Multiple Target Resolution.*** There is always a fundamental limit to the closeness of spacing at which two targets can be resolved, or separately observed as distinct objects, by radar. For frequency-modulated radar, this limit is set by the practical necessity of limiting the width of the band over which the transmitted frequency is swept in modulation. So far as the circuits used in practical indicators are concerned, the resolution limit appears as an effect of the limited speed of circuit response, taken in relation to the limited time available to modulate the transmitted frequency across the entire band used.

The beat-signal phase jump which occurs as indicated in Fig. II.-5(b) at the end of each sweep of transmitted frequency prevents carrying information over from one sweep to the next. Multiple targets must therefore be distinguishable by means of the beat signal received during a single sweep if they are to be resolved at all. Each target will give during one sweep, in accordance with its range R and the width of sweep $F_2 - F_1$ or W, ΔN beat cycles as indicated by (II.24). In order that two targets may be definitely distinguished, the respective numbers ΔN of beat-signal cycles per sweep to which they give rise must differ by a significant amount.

If a minimum difference of $\delta(\Delta N)$ beat cycles per sweep permits a definite distinction to be made between two signals present simultaneously, inversion of (II.24)

gives

$$\delta R = \tfrac{1}{2}(c/W)\,\delta(\Delta N) \qquad \text{(II.26)}$$

as the minimum range separation δR at which two targets may be resolved as distinct objects. This minimum resolved separation varies inversely with modulation sweep width W and is independent of both radio carrier frequency F_0 and modulation frequency f_m. The quantity c/W may be conveniently referred to as *sweep wave length* λ_w. Minimum separation δR for multiple-target resolution is proportional to sweep wavelength by the factor $\tfrac{1}{2}\delta(\Delta N)$.

The exact difference $\delta(\Delta N)$ in beat cycles that is definitely observable depends on the apparatus and technique by which the radar data is indicated and observed. It is fairly obvious, however, that two sinusoidal signals present simultaneously in a single circuit for only ΔN cycles will not be clearly distinguishable as separate if their respective values of ΔN differ by only a very small fraction of a cycle. It is also fairly obvious that the fact of the presence of two distinct signals will be easily discernible if, during a limited time interval in which both signals are simultaneously present, the numbers of cycles ΔN executed respectively by them differ by a considerable number of whole cycles.

Between the above clear-cut limiting cases, it is evident that the somewhat indefinite border line for target resolution occurs for a value of $\delta(\Delta N)$ in the general neighborhood of one complete beat cycle per modulation sweep. That is, targets separated by the order of one-half sweep wavelength will be resolved, with the limit determined more exactly by details of equipment and skill of observer. Resolution is further discussed for more specific conditions in Chapter IX.

A similar limit on resolution in the case of pulse radar is set by sharpness of pulse. This varies inversely with width ΔF of radio-frequency band occupied by the pulse transmission and is, in terms of range, proportional to $c/\Delta F$. That is, either pulse or f-m radar will resolve targets separated in range by an amount δR directly proportional to velocity of radio-wave propagation and inversely proportional to width of radio-frequency channel

utilized. The exact value of the constant of proportionality depends in each case on specific conditions of use, but is of the same order of magnitude whether the transmitted radar signal is pulsed or frequency modulated.

5. WORKING RANGE

a. *Signal Strength on Target.* Strength of received signal is controlled by the same factors in f-m radar as in the better known case of pulse radar, but some discussion will be included here. More on this subject may be found in the literature of pulse radar.[2, 3, 4]

If an antenna in unobstructed space radiates a total power P_t uniformly in all directions, then a power $P_t/(4\pi R^2)$ will fall upon any unit area located at radius R from the radiator and oriented perpendicularly to the direction of the transmitter. This is because the total radiated power P_t reaches altogether at radius R a sphere of area $4\pi R^2$. If the transmitting antenna is directive, some areas will receive relatively more energy at the expense of others which receive less.

The power gain G of a directive antenna in any particular direction is the ratio of the power required by a completely non-directive antenna (at the same location as the directive antenna) to the power P_t fed to the directive antenna, in order that the two antennas may transmit the same signal power per unit area in the given direction. For a simple dipole in the direction of maximum radiation, for example, G is $3/2$. The power received by unit area at distance R from a directive transmitting antenna fed with power P_t and having power gain G_t toward that area is

$$p = P_t\, G_t/(4\pi R^2). \qquad \text{(II.27)}$$

Subject to some qualification with regard to radiator shape, the power gain on the axis of a uniform plane-wave broadside radiator of area A_t is

$$G = 4\pi A_t/\lambda^2 \qquad \text{(II.28)}$$

for wave length λ. Conversely, where power gain can be directly measured but actual radiator area is rather indefinite, this relation defines an "effective radiator

area". A completely non-directional radiator thus has an effective area of $1/(4\pi)$ or 0.0795 square wave lengths, for example, independently of its actual size. Incident power per unit area delivered on a target at range R becomes

$$p = P_t A_t/(\lambda^2 R^2). \tag{II.29}$$

b. ***Target Area.*** Any radar target may be characterized by an "effective echoing area". This is the area which would have to pick up signal, all subsequently re-radiated or scattered uniformly in all directions, in order to produce the same reflected signal in a given direction as does the actual target. Real targets are of complex shape and their effective areas usually vary in a complicated way with direction and wave length, and correspond only in a general way to actual size of target. The apparent power P' scattered by a target at range R, as seen from a direction for which the target exhibits an echoing area A_e, is

$$P' = pA_e = P_t A_t A_e/(\lambda^2 R^2). \tag{II.30}$$

No general discussion of target areas will be attempted here, but the characteristics of some ideal limiting cases[5,6] may be mentioned. The effective echoing area of a target (neglecting any energy dissipation in the target) will be the area intercepting signal from the radar transmitter times the directive power gain, in the direction of the radar receiver, of the target as a radiator. For a conducting sphere very large compared to the wave length, the echoing area is found to be just equal to the actual area of wave front intercepted, which is the projected area of the sphere. For a very large flat conducting plate normal to the line of sight, the echoing area expressed in square wave lengths is proportional to the square of the actual area of the plate, also in square wave lengths. For a very small conducting sphere or flat plate, the echoing area in square wave lengths is proportional to the cube of the actual area in square wave lengths. For an unloaded (parasitic) half-wave resonant dipole, maximum echoing area is 0.86 square wave length. Even with wave lengths as great as a few feet, a medium-sized ship seen broadside can exhibit an effective radar

echoing area of the order of 100,000 square feet.

c. ***Received Signal.*** The signal power per unit area reaching a receiver adjacent to the transmitter, after scattering by a target at range R with apparent total power P', is

$$p' = P'/(4\pi R^2) = P_t A_t A_e/(4\pi\lambda^2 R^4). \qquad \text{(II.31)}$$

A receiving antenna will deliver to a matched-impedance load the power falling upon its effective area A_r to give a total received power

$$P_r = p' A_r = P_t A_t A_e A_r/(4\pi\lambda^2 R^4). \qquad \text{(II.32)}$$

Either (II.28) or (II.32) may be taken as defining A_r; both lead to the same value. In radar work, transmitting and receiving is usually done either with a single antenna or with two identical antennas, so that the same effective antenna area A_a represents both A_t and A_r. It is this identity of transmitting and receiving-antenna characteristics that has made it permissible to disregard wave polarization in the above discussion of power levels.

Equation (II.32) works properly with areas, wave length and range expressed in any mutually consistent units. However, it is especially convenient to express range in wave lengths as length units and all areas in square wave lengths. Then

$$P_r/P_t = A_a'^2 A_e'/(4\pi R'^4), \qquad \text{(II.33)}$$

where primes indicate measurement in wave-length units. This relation is so simple, important and useful that it is well worth remembering.

So long as all dimensions remain the same on a wave-length scale, received power is seen not to depend on wave length used. The target is usually not under control of the radar designer, however, so the possible dependence on wave length of its scattering area in square wave lengths must not be overlooked. As might be expected, the received power varies proportionately with the transmitted power, but a large change in transmitter power has no more effect than a much smaller change in range. Changing the size

(linear extent) of the antennas has as much effect as changing range by the same factor.

d. *Effect of Ground.* When transmitted signal can reach the receiver both by reflection from the target only, and indirectly with intermediate reflection from ground (or sea) as well, the received-signal power varies from practically zero to practically sixteen times the direct-ray value, depending on relative positions of radar, target and ground. This is, of course, due to addition with varying phase of the signals over four paths (two outgoing, two incoming) as the relative path lengths vary with target position. The effect is to divide the space where the target may be into regions in which the target can be seen well by radar and regions in which it can not be seen at all.

For a single radio frequency and a radar location fixed with respect to ground, these regions are fixed in space with respect to the radar. As the radio frequency is changed by frequency modulation, this interference pattern of regions of good and bad radar seeing moves in space. The consequence is that, in the presence of ground reflection, the strength of signal received after reflection from a target varies over the frequency-modulation cycle. That is, the radio-transmission medium becomes frequency selective when multiple paths are present, and produces amplitude modulation of the frequency-modulated signal.

In the general case it is necessary to take account separately of the phase and amplitude, as determined by path lengths and the electrical constants of the ground in the region of reflection, of the signals arriving at the receiver directly and by one or two ground reflections. The total radar signal results from combination of four such components. The properties of the general case may become rather complex, but there are two limiting cases for each of which the results may be presented in a simple form peculiar to the case. In the case of ground radar operating against airborne targets, reflection occurs primarily in the vicinity of the radar and the effect of ground may be treated as an alteration of the directional pattern of power gain or effective area of the antenna used. In the use of airborne radar against surface targets, a typical

f-m application, reflection occurs primarily in the vicinity of the target and the effect of the earth's surface may conveniently be treated as a modification of the directive pattern of effective area of the target only.

For radar altimeters, the surface of the earth is itself the intended target and quite a different condition prevails. Various special cases arise, according to the particular properties of that part of the surface "illuminated" by the radar transmission. So long as the transmitting antenna has even moderate downward directivity, however, practically all the power transmitted must strike the target and the total power scattered or reflected in generally upward directions will depend only on the power transmitted and on the effective reflection coefficient of the surface illuminated.

A very smooth surface of sea water, for example, will act as a mirror with practically perfect reflection capability, giving the effect of direct radiation from an image of the transmitting antenna to the receiving antenna, over a distance of twice the altitude. For an altitude of R' wave lengths and a common effective antenna area of A_a' square wave lengths, the ratio of received to transmitted power will then be[7]

$$P_r/P_t = A_a'^2/(2R')^2. \qquad \text{(II.34)}$$

If the ground surface is so rough as to be a perfectly diffuse reflector, the scattered radiation will be nondirective over the entire upward hemisphere. If the transmitter is highly directive, so that no correction need be made for obliquity of wave travel, the ratio of received to transmitted power for the diffuse case becomes

$$P_r/P_t = \rho A_a'/(2\pi R'^2), \qquad \text{(II.35)}$$

where ρ is the average energy-reflection coefficient for radio waves of the ground area involved.

Equation (II.34) sets an upper limit to power that can be received, except in cases of focusing by special ground configurations, but antenna directivity in altimeters is seldom sufficient to render (II.35) accurately applicable as a lower limit. Dependence of received power on only

the inverse square of the altitude should be particularly noted, in contrast to the inverse-fourth-power dependence (equation II.33) found for targets of fixed area. For nearly horizontal transmission, the effect of ground reflection (perhaps expressed in terms of target directivity) is to produce an inverse-eighth-power dependence of received power on range of a fixed-area target.

e. *Noise.* Beside providing signal pickup, an antenna also acts as a source of random noise of available power $kT\Delta f$ watts,[8] where k is the Boltzmann gas constant, 1.37×10^{-23} watt-seconds per degree Centigrade, T is the effective absolute temperature in Centigrade degrees of the space seen by the antenna, and Δf is the noise band width of the system used in cycles per second. The receiver always contributes additional noise energy, which may be specified by a "noise factor" $\overline{NF}$. This noise factor may be so defined that the output of the actual receiver fed by an antenna at room temperature T_0 is the same as would be produced by a perfect, noise-free receiver of the same gain fed by an antenna producing a noise power of $\overline{NF}kT_0\,\Delta f$ watts.

f. *Noise Limitation on Range.* In some cases, radar range is limited by random noise produced by the antenna and receiver. A great deal of argument is possible as to the exact condition under which the noise becomes harmful, but the limit of range will here be considered to occur when the received-signal power is equal to the noise power at the receiver output. Then from equation (II.33) and the above value for noise power, the limiting range in wave lengths is

$$R' = \sqrt[4]{P_t A_a'^2 A_e' / (4\pi \overline{NF} k T_0 \Delta f)}. \qquad \text{(II.36)}$$

This is sometimes called the Radar Equation. It shows that noise-limited range can be improved equally by increasing transmitted power or by decreasing receiver noise factor, but that the improvement is painfully slow in either case. Wave length alters range only as it alters target echoing area A_e' in square wave lengths. The only really effective method of increasing range which is indicated is by increasing the linear dimensions of

transmitting and receiving antennas.

Decrease of working band width is shown by equation (II.36) to be as effective as increase of transmitter power in improving noise-limited radar operating range. With f-m radar, the noise band width is essentially the pass band of the amplifier for the beat-note output and may be quite narrow. Thus, the steady power transmitted in f-m radar may be much smaller than the peak power of a wide-band pulse radar of equivalent range. In fact, both systems should require substantially the same *average* power for the same signal/noise performance.

g. ***Other Limiting Factors.*** In many cases, f-m radar range is limited by one of two factors other than receiver noise. One of these limitations is *feed through,* or signal passing directly from transmitting to receiving antenna. If this is strong, it interferes with proper receiver operation in the usual f-m case of simultaneous transmission and reception. If a feed-through signal is modulated by motion of conductors near the antennas, it can produce strong spurious variations in the total received signal. Obviously, increasing transmitter power increases both feed-through and desired signals in the same proportion, so gives no increase in useful range when feed through is the limiting factor. Feed-through trouble can be minimized by careful attention to antenna design and location, and by certain artifices, but remains an important practical limitation.

The other major range limitation is reflection from the surroundings of the target—"ground clutter" or "sea return". This, also, increases along with desired signal as transmitted power is increased, so that there is no advantage in using more power than is required to raise the sea-return level well above the receiver noise. Increased sharpness of transmitted beam will improve the ratio of signal to sea return by concentrating a larger proportion of the transmitted power on the desired target, but sea return also remains an important limitation on useful range, especially for single-target systems.

Microphonic noise is always a major source of trouble in airborne equipment, but by careful attention to design it

can be kept within bounds. Microphonic noise is not a fundamental limitation in the same sense as are thermal agitation or shot noise, feed through, and sea return.

6. NOTATION AND REFERENCES

a. *Notation.* The following algebraic notation has appeared in this chapter. Listing is in alphabetic order.

A_a	Effective area of antenna for a given direction.
A'_a	Effective antenna area in square wave-lengths.
A_e	Effective echoing area of target for a given direction, or area of equivalent isotropic scatterer.
A'_e	Effective target echoing area in square wave lengths.
A_r, A_t	Effective area of receiving or transmitting antenna respectively.
c	Velocity of radio-signal propagation, 983.24 feet per microsecond in normal sea-level air.
e_1, e_2, e_3	Signal voltage vectors.
f	Frequency of beat between transmitted signal and signal received after target reflection.
f_m	Frequency of periodic modulation of frequency of radar signal.
f_R	Frequency of f-m radar beat due to target range only.
f_S	Frequency of radar beat due only to relative speed of target.
f_u, f_d	Beat frequency during upward and downward modulation sweep, respectively, of transmitted frequency.
Δf	Equivalent noise band width of receiver.
F	Radio frequency.
F_1, F_2	Radio frequencies transmitted at particular instants.
F', F''	Radio frequencies modified by time lag or target motion
$\dot{F}$	Rate of change of radio frequency with time.
ΔF	Width of radio-frequency channel for pulse transmission.
G	Power gain of antenna for a given direction.
k	Boltzmann's gas constant, 1.37×10^{-23} watt-seconds per degree Centigrade.
k_R	Range sensitivity of frequency-modulated radar, in beat-frequency cycles per second per unit speed (for example, per foot).
k_S	Speed sensitivity of frequency-modulated radar, in beat-frequency cycles per second per unit speed (for example, per foot per second).

N	Number of standing waves between radar and target.
N_1, N_2	Numbers of standing waves under particular conditions.
$\dot{N}$	Rate of change with time of number of standing waves between radar and target.
$\overline{NF}$	Noise figure of receiver.
ΔN	Change in number of standing waves caused by frequency change or relative target motion.
$\delta(\Delta N)$	Difference of standing-wave changes for two targets.
p	Power falling on unit area at a distance from radar transmitter.
p'	Power falling on unit area at a distance from reflecting target.
P'	Power reflected by target.
P_r	Available power delivered to matched load by receiving antenna.
P_t	Power radiated by transmitting antenna.
R	Range or distance of target from radar.
R'	Range in wave lengths.
$\dot{R}$	Rate of change of range with time.
δR	Range difference between two targets just resolved by radar.
t	Time.
t_1	Particular instant of time.
t_m	Period of repetition of periodic frequency modulation.
T	Absolute temperature of region seen by antenna, in Centigrade degrees.
T_o	Absolute room temperature in Centigrade degrees.
W	Width of frequency band swept in modulation of transmitted radar signal.
λ	Wave length of radio signal.
λ_W	Wave length of radio signal of frequency W, called sweep wave length.
ρ	Coefficient of radio-wave reflection.
τ	Time delay while signal travels from radar to target and echo returns.
ψ	Phase of reflected signal from target relative to direct signal, at receiver.
$\Delta\psi$	Change of relative reflected-signal phase during modulation.

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CHAPTER III.

RADIO APPARATUS USED IN F-M RADAR SYSTEMS

1. GENERAL

The radio portions of a frequency-modulated radar system include directive antennas for transmission and reception, oscillators for generating the radio-frequency power transmitted, frequency modulators controlling these oscillators, sources of modulating signal, receivers for the target-reflected signals which derive therefrom beat-frequency signals representative of target range and speed, and beat-frequency amplifiers of suitable characteristics. These various portions of the equipment will now be discussed separately, with emphasis on those features which are peculiar to f-m radar use and as little as possible said about their properties as conventional radio apparatus. For airborne use, it is imperative that microphonic properties be minimized in all portions of the equipment.

2. ANTENNAS

a. *Types of Antenna.* Antennas used for f-m radar have been of conventional types for the frequencies involved. Mechanical design has been subject to the requirements of airborne use. Separate antennas have in most cases been used for transmission and for reception.

For altimeters operating near 440 megacycles, the simple low-reactance (large-diameter) half-wave dipole shown in Fig. III.-1 is widely used. It is supported from the outer skin of metal aircraft by the quarter-wave stub section of two-conductor open line seen in the photograph, which functions as a rugged radio-frequency insulator. It is fed at the center by a single coaxial line, of which one of the supporting rods forms the outer conductor. The supporting line stub serves also as an unbalanced-to-balanced transformer to make the antenna feed symmetrical and to match the antenna impedance to the 50-ohm characteristic impedance

of the coaxial feed line. The metal skin of the aircraft to which the antenna is mounted, or a large metal mounting plate used in the case of fabric-covered aircraft, serves as a reflecting sheet to make the downward directivity of the system greater than that of a dipole in free space.

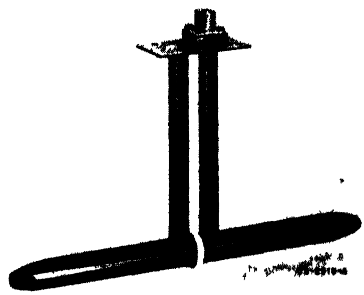

Fig. III.-1. Dipole antenna for 440-megacycle airborne use.

Increasing aircraft speeds have made the aerodynamic drag even of streamlined antennas, like the dipole of Fig. III.-1, excessive. Much effort has therefore been applied to a program for development of antennas completely enclosed within the normal aircraft structure. One result of this program has been the "slot antenna"[1] for 440-megacycle altimeters shown in Fig. III.-2. This unit is to be mounted with its face flush with the metallic outer covering

Fig. III.-2. Slot antenna for 440-megacycle operation. For flush mounting in surface of aircraft.

or skin of an aircraft. It will excite in the skin of the aircraft currents in the plane of and perpendicular to the long slot visible in its face. These currents will radiate substantially like a broad-band dipole antenna, so that the unit functions essentially as a coupling device between a coaxial transmission line from the radar and the skin of

the aircraft. The metal "bath tub" of the unit acts with the three sets of capacitive tabs seen at the edges of the slot as a resonant circuit coupled to the adjacent aircraft skin by the slot capacitance common to both. The transmission line from the radar is in turn coupled to the resonant circuit by a suitably terminated inductive loop mounted within the bath-tub cavity. The open slot is covered by a window of transparent plastic. Substantially the same altimeter performance is obtained with the antenna of Fig. III.-2 as with that of Fig. III.-1.

For higher directivity at 410 megacycles, the Yagi end-fire array of Fig. III.-3 has been used. It is an enlarged version of the 515-megacycle array used with the ASB series of pulse radars. This linear array has one driven dipole

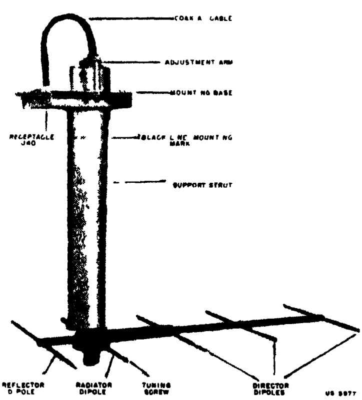

Fig. III.-3. Directional antenna array for use at 410 megacycles per second.

element, located at the end of the streamlined supporting mast, and four parasitic elements, one "reflector" and three "directors", excited by radiation from the driven element. It also is fed by a coaxial 50-ohm line within the supporting mast, through a balancing and impedance-matching stub line within the housing at the end of the mast. The directive pattern of this antenna is primarily a single lobe directed along the bar mounting the director elements and extending between half-amplitude points over a total arc of 76 degrees in the (horizontal) plane of the elements and a total arc of 105 degrees in the (vertical) plane of the supporting rod and mast.

Accurate azimuth determination has been obtained by using in a lobe-switching arrangement[2] two of the Yagi arrays of Fig. III.-3, pointed respectively to right and left of a reference azimuth by a constant small angle (about 15 degrees). When the antennas are turned as a unit until switching between them no longer affects the strength of signal, the target must lie along the above azimuth-reference line, as is well known. Special care in reducing backwardly directed pattern lobes[3] is necessary to avoid azimuth errors in the presence of interfering targets behind the aircraft carrying the antennas.

At 1500 megacycles, the array shown in Fig. III.-4 of two dipoles in a parabolic-cylinder reflector has been used to produce a forwardly directed signal. In keeping with the trend toward higher aircraft speeds and the consequent need to minimize exterior projections, this antenna is intended to be mounted within the aircraft and to radiate through a dielectric window shaped to the contour of a

Fig. III.-4. Dipole array in reflector for 1500-megacycle use, with matching network.

normal exterior surface of the aircraft. The dipoles are mounted to the reflector by resonant stub lines and are connected in parallel to a 50-ohm coaxial cable by means of an adjustable two-stub matching unit. Vertical pattern width is 74 degrees and horizontal width 66 degrees, between half-field points. Some tests have also been made of a two-layer Yagi array[4] and of an end-fire array with fed elements,[5] as well as of larger dipole-parabola

systems.[6] Increased directivity was found to produce improved results. For ship-board use at 1500 megacycles, with relatively high directivity required, a single-dipole antenna with a small reflecting "hat" is mounted in a paraboloidal reflector four feet in diameter.

In exploratory work at 4000 megacycles, a two-dipole array mounted in a paraboloidal reflector of circular form has been used successfully for each antenna. Because space is at a premium on aircraft, use of separate transmitting and receiving antennas is a decided disadvantage and methods of duplexing a single antenna are important. The time-sharing type of duplexing used with pulse radar is not useful with the continuous transmission and reception which is characteristic of most frequency-modulated radar, while bridge methods are not only wasteful of power but excessively critical if extreme decoupling is to be maintained over a wide frequency band. The "Magic Tee" wave-guide coupler is a promising decoupling device for duplex use of a single antenna.

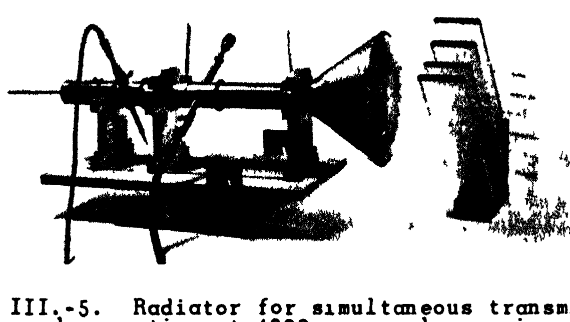

Fig. III.-5. Radiator for simultaneous transmission and reception at 4000 megacycles, using polarization duplexing.

These considerations led to tests of the less conventional 4000-megacycle transmitting and receiving antenna[7] of Fig. III.-5, in which a single horn-and-lens radiator is coupled independently to two circuits by use of two distinct conditions of wave polarization. The horn is fed by a circular wave guide operated in the TE_{11} mode and the different polarizations are obtained by crossing the two line elements coupling the transmitter and receiver to the wave guide, as may be seen in the figure. Plane grids of wires shaped to conform with electric-field lines of the

TE_{11} wave from the rear coupling (the one farthest from the horn) are inserted in the guide between the couplings. These serve to prevent waves from the front coupling line from reaching the rear coupling, as well as to terminate the portion of wave guide seen by the front coupling.

Limited tests of the duplexed antenna gave promising results. A few targets tried were found to scatter radiation in unpolarized fashion, so that returned-signal levels were about the same for crossed as for parallel polarization of separate transmitting and receiving antennas. For targets which reflect radio waves without change of polarization, the quarter-wave plates of parallel dielectric slabs or metal sheets, shown in front of the horn, may be used (when oriented at 45° to both transmitting and receiving couplings). Such devices convert the plane-polarized waves leaving the horn to circularly polarized waves for illuminating the target, and reconvert circularly polarized reflected signals to waves plane polarized at 90° to the outgoing ones for effective reception. Good isolation of transmission and reception was obtained with this duplexed antenna, especially when used without the quarter-wave plate.

b. *Location on Aircraft.* Proper location of antennas on aircraft is very important for satisfactory operation of airborne frequency-modulated radar equipment. Six major requirements to be met in locating the antennas are:

(1) Minimum disturbance of flight characteristics of aircraft.

(2) Minimum coupling between transmitting and receiving antennas.

(3) Minimum coupling to antennas of other equipment.

(4) Minimum modulation of signals by motion of propellers or other portions of the aircraft.

(5) Minimum signal transmission or receiving sensitivity in undesired directions.

(6) Maximum signal in desired direction.

These requirements are likely to prove more or less incompatible, so that a compromise location generally gives the best overall result.

For the dipole antennas of Fig. III.-1, mounting so that the axis of the radiator is parallel to the motion of the aircraft is obviously important to minimize aerodynamic drag. Coupling between transmitting and receiving antennas is minimized if they are mounted with axes in line, since each antenna is then located in a null of the directive pattern of the other. In-line mounting with enough separation to prevent undue induction-field coupling usually requires that the antennas be mounted on the under side of the fuselage, which is not always permissible. Where these antennas must be mounted in side-by-side fashion, they must be well separated or shielded from one another if excessive cross coupling or *feed through* between them is to be avoided. This may be done by mounting one antenna under each wing of a high-wing or mid-wing aircraft; the fuselage then acts as a very effective shield.

The Yagi arrays of Fig. III.-3, being intended to work against isolated targets ahead of the aircraft, must be mounted pointing almost exactly forward. Because the radar echo from the desired target may be very weak and the earth represents in this case a tremendous undesired target, extreme care must be taken to minimize downwardly directed transmission and reception if a very strong unwanted "altitude signal" is to be avoided. This suggests mounting the antennas atop the wings, one to each wing, somewhat behind the leading edge,[8,9,10] as shown in Fig. III.-6A.

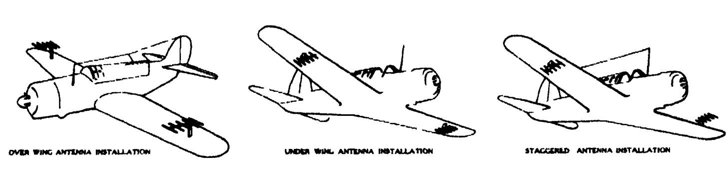

Fig. III.-6. Typical locations for directive antennas on aircraft.

For low-wing or mid-wing aircraft, such mounting permits the fuselage to shield the antennas effectively against direct feed-through coupling, which must be kept extremely small. But with over-wing antennas, the shadow of the wings makes it difficult for the radar to see targets at short range. On the other hand, the under-wing installation of Fig. III.-6B may lead to excessive altitude signal, despite the cancellation of downward signal obtainable by

correct spacing between array and wing surface. The staggered installation shown in Fig. III.-6C, with one antenna above one wing and the other antenna between the other wing, has sometimes proved a useful compromise.

Leading edges of thick wings provide suitable sites for installation of the 1500-megacycle parabolic reflectors of Fig. III.-4.[11,12] With one antenna in each wing, the fuselage is an effective shield preventing either antenna from radiating directly into the other. Downward radiation can be kept low by proper placing of the antenna with respect to the hole cut in the wing. The aircraft propellers must not enter the main radiation pattern of either antenna or serious unwanted modulation will result.

At 4000 megacycles, it has been found feasible to operate both antennas in fixed positions side by side in the plastic nose of a small bombing aircraft. This has required a metallic shielding septum between the antennas and a non-reflecting surface behind them.

When feed-through signal due to direct radiation or induction coupling of transmitting and receiving antennas is substantially eliminated, indirect couplings become troublesome. Such indirect coupling is likely to occur when protruding portions or accessories of the aircraft, such as propellers, landing gear, cowling, antennas, bombs or rockets, pick up and re-radiate a portion of the transmitted signal. Indirect feed through can be extremely troublesome because motion of the offending conductor or variable electrical contact between it and the aircraft produces strong modulation of the feed-through signal, which then becomes interfering noise. Direct feed through merely simulates a stationary target at very short range; this does harm only by confusing desired near-target signals and by producing receiver overloading which may mask weak distant-target signals. Simultaneous transmission and reception makes the feed-through problem more serious for f-m than for pulsed radar. For satisfactory operation, transmitted signals reaching the receiving antenna in the absence of desired targets should suffer at least 60 decibels attenuation.

Cancellation of feed through by fine adjustment of coupling elements is usually of limited value. It may only

be depended upon to reduce further a coupling already inherently small, and this only if antennas are so disposed that no large chance variation of indirect coupling can occur. Highly accurate cancellation can be produced at a single frequency, but only a limited attenuation can be maintained in this way throughout a wide frequency band.

Unwanted echoes from rough sea surfaces nearly beneath the aircraft are reduced by the measures against downward directivity that are used to overcome unwanted altitude signal. At least for the lower frequencies of a few hundred megacycles, sea return is much worse for vertically than for horizontally polarized transmission. Antennas for other than altimeter use must therefore be mounted so that, even in the presence of possible inclined aircraft surfaces,[13] minimum vertical polarization will be produced. Signals returned from rough sea surfaces near a distant target are essentially similar to those from the target; interference from them can only be reduced by transmitting and receiving with narrow beams to reduce the amount of interfering surface illuminated or seen by the radar.

From this brief discussion, it should be evident that antenna placement is an important branch of the art of airborne frequency-modulated radar. Unfortunately, the fields in the immediate vicinity of an aircraft are so complex that detailed theoretical prediction of antenna behavior seems impracticable. Each new combination of antenna and aircraft requires individual study, preferably involving directive-pattern measurement and actual radar-system flight tests for several promising antenna locations. A general understanding of the behavior of radio-wave fields and a background of experience are the only useful guides in predicting optimum locations for antennas on aircraft.

3. TRANSMITTING OSCILLATORS

a. *Triodes.* Self-excited oscillators used to generate frequency-modulated radar signals have, like antennas, been of conventional design for the frequencies used. Triode tubes have therefore been used for the lower frequencies. The push-pull resonant-line oscillator using acorn tubes which is shown pictorially in Fig. III.-7 and schematically in Fig. III.-8 is typical of 400- to 500-megacycle design.

This oscillator develops about 0.20 watt and is tuned to the proper average frequency by sliding a short-circuiting bar along the plate line. A rather similar arrangement using type *2C43* lighthouse triodes provides a 2-watt power output, with plate current supplied at 270 volts.

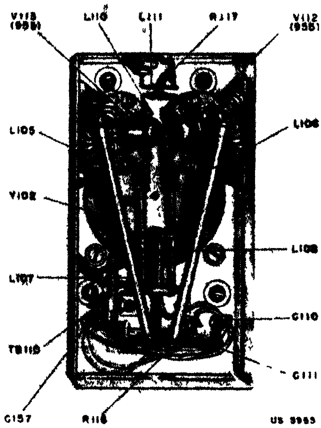

Fig. III.-7. Frequency-modulated triode oscillator for 410 megacycles.

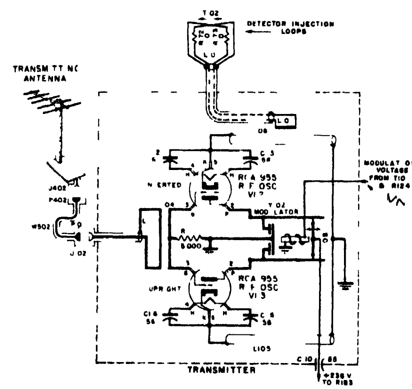

Fig. III.-8. Circuit diagram for 410-megacycle oscillator.

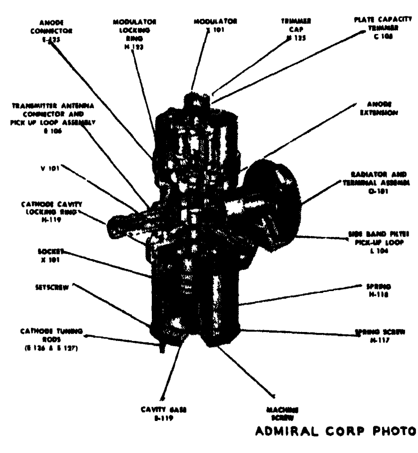

Fig. III.-9. Triode oscillator for 1500 megacycles.

At 1500 to 1650 megacycles, an oscillator using a single *2C43* tube with grounded grid in a tuned-plate tuned-cathode coaxial structure has proved suitable. As arranged in

Fig. III.-9, with the half-wave-length plate line supported at its nodal point by a quarter-wave stub of high heat conductivity, this unit will deliver up to 2 watts of r-f power, again with 270-volt plate supply. A coupling link between plate and cathode resonators, not visible in the figure, provides adequate feed back to ensure oscillation despite the shielding action of the grounded grid. Tuning is accomplished by adjustment of capacity loading of the open-ended half-wave plate line, while loading of the tube is controlled by varying the length of the cathode line in the region above the resonant length.

Each type of triode oscillator described above is frequency modulated by means of a tuning capacitor with a vibrating electrode at ground potential. The oscillator is tuned and loaded so as to minimize accompanying amplitude modulation. In the lower-frequency push-pull oscillator, the modulating capacitor has symmetrical fixed electrodes connected to the resonant line at the tube plates. In the single-ended oscillator, the single fixed electrode of the modulating capacitor forms the end of the half-wave resonant line remote from the tube plate. Total frequency swings ranging from 1 to 50 megacycles per second are used.

b. ***Magnetrons.*** At 2600 megacycles, triodes do not make particularly satisfactory oscillators. To provide about one watt of frequency-modulated continuous-wave power at that frequency, which was considered for f-m radar use, a number of forms of special magnetron with a single annular resonant cavity were tried.[14] These tubes were frequency modulated by varying anode voltage. In one experimental tube, an efficiency of 25 per cent was attained with 200 volts on the anode, and in another a 25-megacycle frequency swing was obtained. Tubes of this type are therefore usable and might respond to further development, but none of those tried was outstandingly good.

At 4000 megacycles, the multicavity magnetron is the most efficient transmitting oscillator. A small 12-cavity continuous-wave magnetron was successfully developed for 4000-megacycle f-m radar tests, under the experimental type designation *A-125C*. A modification of this tube with electronic frequency modulation of a type to be described

later was also developed and used with success. Either tube delivers 25 watts of r-f power to conventional coaxial load circuits at about 50 per cent efficiency, with d-c plate supply at 850 volts.

Under some conditions of operation, these magnetrons were extremely noisy, exhibiting strong random modulation of both amplitude and frequency of their r-f output. This was found[15] to be caused by liberation in the cathode-anode space of a condensible vapor of atoms and ions of the cathode-coating material, as a result of the back bombardment of the cathode by electrons accelerated by the radio-frequency fields in the oscillating tube. Excess noise occurs only when operating at voltages above a certain threshold value, which depends on the structure and past history of the individual tube and on the current being drawn. Any measure which reduces electronic bombardment of the cathode raises the noise-threshold voltage, as well as improving the operating life and efficiency of the tube. Such measures are omission of oxide coating on the ends of the cathode cylinder, which operate in regions of non-uniform field, and use of cathode sleeves of star-shaped external cross section. With oxide coating only on the clockwise faces of the star, clockwise electron rotation ensures that only the uncoated faces are bombarded. When the noise-threshold voltage of a tube exceeds its normal operating voltage, the tube is satisfactory from a noise standpoint for f-m radar use.

4. MODULATORS

a. *Rotary Capacitors.* Early altimeters[16] were modulated in frequency by varying the transmitter tuning rapidly with the aid of a motor-driven rotary variable capacitor. This is a simple and obvious method, which can within reasonable limits be given any desired form of frequency variation with time by choice of a suitable electrode shape, but suffers from several practical disadvantages. It was shown in Chapter II., section 4g, that the range sensitivity of a periodically frequency-modulated radar is proportional to the product of the repetition frequency of the modulation and the width of the frequency band swept as a result of the modulation. It is difficult to maintain the speed of an ordinary modulator-driving motor, and

consequently the modulation frequency of a rotary capacitor, sufficiently accurately over the wide ranges of temperature and supply voltage normally encountered in airborne operation. It is also difficult to set the width of the band swept to desired values and to change from one value to another for operation in different altitude ranges. Mechanical play results in unintentional frequency modulation, which produces a "noisy" signal.

An ingenious scheme was used in the rotary modulator of the German radar altimeter *FuG-101,* to alter the band swept without changing modulation frequency. This employed a capacitor with one large and one small rotor. The small rotor was driven by the motor directly and the large rotor by a unidirectional clutch. With the motor running in one direction, both sections were driven to sweep a wide frequency band, while reversal of the motor disengaged the unidirectional drive and operated only the small rotor element.

b. *Vibrating Capacitors.* The type of frequency modulator that has seen most radar use is one in which an element of the capacitor which tunes the transmitting oscillator is driven by a reciprocating electric motor and executes a linear vibratory motion. The motors used are essentially electrodynamic loudspeaker movements, driven synchronously from an alternating-current source of controlled frequency. Synchronous a-c drive is more practical for vibratory than for rotary modulators because of the smaller driving power required by the former.

Fig. III.-10 shows the modulator used with the 1500-megacycle oscillator described in section 3a above. The larger unit on the left contains the permanent-magnet field structure. In the recess of its face may be seen the perforated diaphragn which supports the actuating "voice" coil and the cylindrical moving electrode of the modulating capacitor. Within the vibrating cylinder is visible a flat disc electrode, which is adjustable by a screw passing through the axis of the structure and accessible from the rear; this disc serves for fine tuning of the transmitter to the assigned operating frequency. The smaller unit at the right of the picture is the mechanically fixed and electrically insulated cylindrical electrode of the modulating

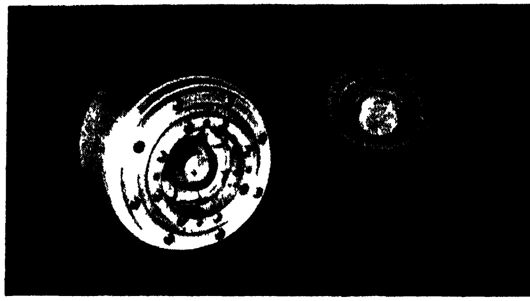

Fig. III.-10. Vibrating capacitor for modulating transmitter frequency.

capacitor, which enters the vibrating cylinder and faces the adjustable disc of the driving unit when the two pieces are united to form a complete modulator assembly. Because the modulating capacitor has coaxial cylindrical electrodes with relative motion in the axial direction only, its capacitance varies in accurately linear fashion with displacement of the moving electrode.

The modulator used with the 410-megacycle transmitter described in section 3a is similar in principle to the one illustrated above. Its moving electrode is, however, a flat central portion of the diaphragm and its symmetrical fixed electrodes are plated areas on a flat ceramic head plate mounted close and parallel to the diaphragm. This parallel-plate type of capacitor with variable spacing exhibits an inherently non-linear variation of capacitance with diaphragm displacement.

On a lumped-circuit basis, oscillator frequency is well known to be inversely proportional to the square root of the total capacitance in the frequency-determining resonant circuit. If this capacitance is entirely that of a parallel-plate capacitor having negligible field fringing, it will in turn be inversely proportional to the separation of the plates. The oscillator frequency will in that case be directly proportional to the square root of the plate separation, which is varied linearly by motion of the modulator diaphragm. If the capacitance is entirely that of two coaxial cylinders with negligible fringing, it will be directly proportional to the depth to which the smaller cylinder enters the larger one. The oscillator frequency will in this case be inversely proportional to the square

root of the depth of cylinder engagement, which is varied linearly by motion of the diaphragm in the type of modulator shown in the figure. Thus the two types of modulator described both give non-linear variation of frequency with electrode displacement and to a similar degree. Had additional capacitance not varied by modulator motion been taken into account, a similar conclusion would still have been reached.

To realize truly linear variation of frequency with displacement, more complicated electrodes than parallel planes or coaxial cylinders would be required. Fortunately, the simple configurations have sufficed for the relatively limited frequency variations used. The coaxial-cylinder form of modulating capacitor has the property that its rate of change of capacitance with electrode motion is independent of initial depth of engagement of its electrodes. This means that amplitude of frequency modulation is, for a given amplitude of diaphragm vibration, unaffected by deformation of the diaphragm due to acceleration of the modulator as a whole. For use in mobile craft, freedom from disturbance of the frequency swing by acceleration is an important advantage. The parallel-plate modulator does not exhibit this advantage.

Overall sensitivity of the modulator, in amplitude of capacitance change per volt of driving-signal amplitude, directly affects the frequency swing and therefore the range sensitivity of the radar system, so must be stable for a single modulator and tolerably uniform among production modulator units. Sensitivity depends mainly upon electrical resistance and conductor length in the driving coil, magnetic flux density in the air gap containing the coil, and mechanical properties of the moving system. Since the modulator is normally driven at frequencies below its natural resonance, the controlling mechanical property is diaphragm stiffness. These quantities must be held to design values in production, and wherever possible chosen to vary in mutually compensatory fashion with such external factors as temperature. Diaphragm stiffness, for example, may be controlled by varying the thickness and tension of the diaphragm. A threaded ring for tension control, with two slots to receive an adjusting spanner, may be seen

between the two rings of screwheads in Fig. III.-10. Vibrating frequency modulators must be recognized as highly specialized units requiring great care in design and manufacture. Typical values for production parallel-plate modulators are given in Table III.-1.

TABLE III.-1

Properties of Vibrating Modulator

Driving-Coil Resistance	5.7 ohms.
Driving-Coil Inductance	Negligible.
Flux-Density × Conductor Length	4.0×10^6 gauss-cm.
Mechanical Resistance	112 gm./sec.
Diaphragm Stiffness	2.4×10^4 dynes/cm.
Moving Mass	2.0 grams.
Resonant Frequency	173 cycles/sec.
Operating Frequency	120 cycles/sec.
Sensitivity	0.011 inch/volt.

The diaphragm can "break up" and vibrate in variously phased sections at high frequencies. If high-frequency exciting components are present in the driving signal, such break-up can seriously disturb the overall modulation characteristic. By particular care in design, higher-mode resonances of the moving system can be made very weak; this is achieved in the modulator shown in the figure.

c. *Driving the Vibrating Modulator.* For small frequency swings, nonlinearity in the effect of diaphragm motion on transmitter frequency may be neglected. Symmetrical-sawtooth frequency-modulation wave form then requires that the vibrating-modulator diaphragm move with constant speed for one half modulation period, then reverse its motion suddenly and again move with the same constant speed but in the opposite direction for the next half period, then again reverse suddenly, and so on. Curve (a) of Fig. III.-11 represents the triangular variation of diaphragm displacement with time during this motion, curve (b) the corresponding square-wave variation of velocity and curve (c) the impulsive acceleration of the diaphragm.

Deformation of the diaphragm, opposed by its stiffness, requires a driving force proportional to displacement and therefore representable, to the proper scale, by curve (a). Any mechanical resistance of the type corresponding to

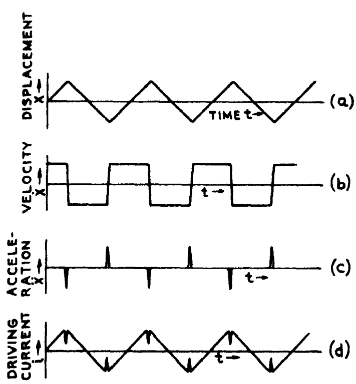

Fig. III.-11. Characteristics of modulator-diaphragm motion.

motion through a viscous fluid must be overcome by a force proportional to velocity and representable by curve (b). For this particular motion, friction also requires a driving force of form (b). Acceleration of the moving mass at the points of reversal requires an inertia force representable by (c).

The total driving force, the sum of these three components in proportions depending on the particular properties of the moving system, must have the sort of time variation shown by curve (d) of the figure. To avoid breaking up of the diaphragm into regions vibrating at high frequency and not in phase, it is necessary in practice to spread out the impulsive driving force and so avoid extremely sudden reversals of motion at the turn-around points. It is also found in practice that the viscous-force component (b) required is negligible by comparison with stiffness and inertia forces (a) and (c).

With the mutually perpendicular arrangement of magnetic field, current, and motion found in the vibrating-modulator motor, the driving force developed is simply the product of magnetic induction in the air gap, conductor length in the driving coil, and current through the coil. The current must therefore also have the wave form of curve (d). The corresponding driving voltage will have a similar form, because the electrical impedance of the moving voice coil itself is practically a pure resistance. Curve (d) therefore represents the wave form of the voltage which must be

applied to the modulator-driving amplifier in order to transmit a frequency varying according to curve (a).

The driving wave form may conveniently be obtained from a square-wave voltage source because the required symmetrical-sawtooth component represents the variation of area under, or mathematically the time integral of, a square wave, while the pulse component represents the rate of change or time derivative of a square wave. The charge on, hence the voltage across, a capacitor is by definition the time integral of the current flowing into that capacitor. Conversely, the current flowing into a capacitor is proportional to the rate of change or time derivative of the voltage applied across that capacitor. A capacitor fed with a square-wave current, for example through a very high resistor from a square-wave voltage, therefore develops a symmetrical-sawtooth voltage. A capacitor across which is impressed a square-wave voltage passes an impulse-wave current, which will produce an impulsive voltage across a very small series resistor. Substantially all of the total applied voltage must appear across the "very large" resistor above and substantially none of it across the "very small" resistor, if accurate integration and differentiation of wave form are respectively to be attained. That is, the time constant of the integrating circuit must be large and that of the differentiating circuit small.

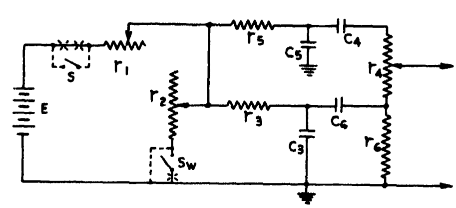

Fig. III.-12. Modulator-driving circuit.

These considerations led to the modulator-driving circuit of Fig. III.-12, which has proved very useful. The initial square wave is produced from a low-impedance constant-voltage supply E by periodic operation of a rapid-acting switch Sw at the desired modulation frequency. A well damped, cam-driven leaf-spring mechanical switch with equal open and closed periods has been used, either as shown by the full-line circuit or by the alternative dotted

circuit. Tests have shown a low-impedance electronic switch[17] to be an acceptable though more complex substitute. The high-level square wave is adjustably attenuated by the moderately low-valued resistors r_1 and r_2, either or both of which may be varied as specific conditions require. These resistors provide convenient control of the frequency band swept in modulation and thereby of the range sensitivity of the radar system.

The output of the attenuator is applied to two wave-shaping circuits, in which the relatively high resistor r_3 and large capacitor C_3 integrate the square wave to produce a triangular-wave output, while the relatively small capacitor C_4 and small resistor r_4 differentiate to provide pulse output. Resistor r_5 and capacitor C_5 act to decrease the steepness of rise and fall of the square wave before differentiation, preventing the derivative pulses from being so sharp as to cause marked breaking up of the piston motion of the modulator diaphragm. Large capacitor C_6 and resistor r_6 serve merely to block any steady voltage from appearing at the output, while maintaining a closed output circuit for direct current.

Connection of the bottom of pulse-output resistor r_4 essentially to the top of sawtooth-output capacitor C_3 results in direct addition of the integrated and differentiated waves, in proportions adjustable bv the variable tap on r_4 to suit the mass/stiffness ratio of the modulator. Low-level output of appropriate wave form from the tap on r_4 drives an amplifier with a low-impedance, or constant-voltage, output circuit which in turn drives the modulator.

Resistor r_5 is usually large enough and capacitor C_5 small enough not to load the attenuator r_1, r_2. Resistor r_3 may be small enough, however, to produce appreciable loading and so affect the calibration of the attenuator. Capacitor C_3 is large enough so that the voltage across it remains nearly constant at the mean value of the attenuator output over the square-wave cycle, since the triangular-wave voltage cycle across C_3 is of relatively small amplitude. r_3 is therefore connected in effect from the square-wave attenuator-output voltage to a point at a fixed voltage having the same average value. Its effect is therefore the same as that of a shunt having substantially

twice the resistance of r_3 and connected directly across r_2. Allowance for loading by r_3 may thus be made easily in the attenuator design.

Radar range sensitivity is controlled by the rate of change of transmitted frequency with time during modulation. This rate must therefore be accurately controlled if accurate range measurement is required. In the circuit of Fig. III.-12, the current flowing into capacitor C_8, to which the rate of change of capacitor voltage, of diaphragm displacement, and of transmitted frequency are proportional, depends only on supply voltage E, settings of attenuator resistors r_1 and r_2, and series resistance r_3. This important rate is therefore not dependent on modulation frequency in the case of the circuit shown. Fig. III.-13 is a graphic demonstration of the way in which current

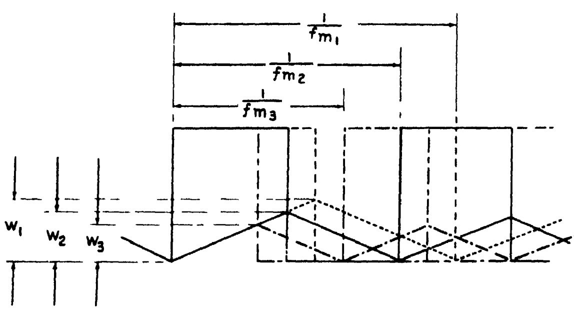

Fig. III.-13. Automatic compensation of modulation-frequency changes by altered sweep width.

integration varies the width W of band swept so as to compensate for variation of modulating frequency f_m, keeping constant the modulation product Wf_m which is one half of the rate of change of transmitted frequency $\dot{F}$. So long as the proportion of triangular to pulse component does not depart excessively from the stiffness/mass ratio of the modulator as modulating frequency varies, the circuit shown compensates radar range sensitivity against such variation.

The entire modulating system is one of the few portions of a complete f-m radar equipment which can affect directly the accuracy of its operation. Careful design and accurate, stable components are therefore necessary throughout this portion of the equipment.

d. *Electronic Modulators.* Vacuum tubes with feed back so connected as to give the tube the properties of a variable reactance, often used as oscillator-frequency controls at medium and high frequencies, have not found use in ultra-high-frequency frequency-modulated radar. Their range of control, stability and control linearity do not appear adequate for such use, even in the case of tubes having low enough electron-transit time to be operative at all at ultra-high frequencies.

A Barkhausen transit-time oscillator with frequency control by variation of accelerating voltage was used in an early altimeter[18] but was probably not very accurate. In similar fashion, a reflex klystron with frequency modulated by varying repeller voltage has been given limited trial but, aside from inadequate power output, exhibited inadequate stability and linearity of modulation for accurate range measurement. Successful use of such an oscillator in a special superheterodyne receiver will be described later.

Most of the work on electronic frequency modulation of radar signals has been done in connection with magnetron power oscillators operating at super-high frequencies in the neighborhood of 4000 megacycles per second. Requirements to be met by the modulator are:

(1) Sweep capability not less than 5 megacycles per second.

(2) Minimal accompanying amplitude modulation.

(3) Linear frequency-modulation characteristic.

(4) Modulation characteristic independent of oscillator load.

Direct modulation of magnetron anode voltage does produce frequency modulation, but only with excessive accompanying amplitude modulation. Since the resonant-cavity circuit controlling the oscillating frequency is self-contained within the magnetron envelope, direct variation of a main tuning reactance by the methods used at lower frequencies is inconvenient. Frequency control by external circuits coupled to the magnetron has therefore been extensively investigated.[19] The rather low coupling attainable between any external circuit and the resonant cavities of magnetrons that were available during the f-m

radar work has been found to impose practical limitations.

A magnetron with a resonant cavity coupled to it by the usual coupling loop may be represented well enough for the present discussion by the simple equivalent circuit of Fig. III.-14. This, of course, assumes that

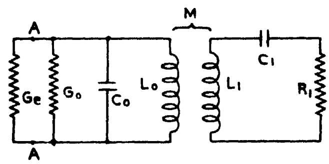

Fig. III.-14. Equivalent circuit for external frequency modulation of magnetron oscillator.

no other modes of resonance of the magnetron cavities or the coupled cavity than those so represented lie in or very near the frequency range under consideration. G_0 represents losses occurring in the magnetron cavities, including any separately coupled external load, in the absence of the coupled circuit under investigation. G_e is the conductance of the electron stream of the magnetron and is found to vary very slowly with frequency; any susceptance due to the electron stream is included in the lumped-equivalent magnetron inductance L_0 and capacitance C_0.

For the present purpose, the behavior of the equivalent circuit is fully described by the magnetron oscillation frequency f_0 in the absence of the coupled circuit, given by $1/(2\pi\sqrt{L_0C_0})$, and its Q-factor ω_0C_0/G_0, in addition to the coefficient of coupling k, which is $M/\sqrt{L_0L_1}$, and the Q-factor $1/(\omega_1C_1r_1)$ of the external circuit at a frequency f_1 given by $1/[2\pi\sqrt{L_1C_1(1-k^2)}]$, which characterizes the effect of adding the external coupled circuit of uncoupled resonant frequency $1/(2\pi\sqrt{L_1C_1})$. Uncoupled frequencies, uncoupled Q and coefficient of coupling are all measurable characteristics of the resonant system, while the lumped equivalent impedance elements L_0, C_0, M, L_1, C_1 and r_1 are not.

Susceptance B appearing across terminals A-A of the equivalent network at frequency f is given by

$$B = G_0 Q_0 \left\{ f/f_0 - f_0/f + \frac{k^2}{1-k^2} \cdot \frac{Q_1^2 \left[f_1/f - f/f_1 \right] f_0/f_1}{1 + Q_1^2 \left[f_1/f - f/f_1 \right]^2} \right\} \qquad \text{(III.1)}$$

and conductance G at the same points by

$$G = G_0 \left\{ 1 + \frac{k^2}{1-k^2} \cdot \frac{Q_0 Q_1 f_0/f_1}{1 + Q_1^2 \left\{ f_1/f - f/f_1 \right\}^2} \right\} \qquad \text{(III.2)}$$

Oscillation of the coupled systems must take place at a frequency for which susceptance B is zero. Only external-circuit resonant frequencies f_1 and coupled oscillation frequencies f which lie very near to the uncoupled magnetron frequency f_0 are of present interest. Fractional frequency deviations

$$\left.\begin{aligned} y &= (f - f_0)/f_0 \\ \text{and} \qquad x &= (f_1 - f_0)/f_0 \end{aligned}\right\} \qquad \text{(III.3)}$$

are therefore useful variables; these remain so small that they may be neglected by comparison with unity. Using these variables, the condition for vanishing B which determines the frequency of oscillation becomes

$$y = \frac{k^2}{1-k^2} \cdot \frac{Q_1^2 (y-x)}{1 + 4\, Q_1^2 (y-x)^2} \qquad \text{(III.4)}$$

to a good approximation, while the conductance becomes

$$G = G_0 \left\{ 1 + \frac{k^2}{1-k^2} \cdot \frac{Q_0 Q_1}{1 + 4Q_1^2 (y-x)^2} \right\} . \qquad \text{(III.5)}$$

From equation (III.4) it is clear that the change $f_0 y$ in oscillation frequency caused by coupling the external circuit to the magnetron depends only upon the degree of coupling, upon the selectivity of the coupled circuit, and upon the detuning $f_0 x$ of the coupled circuit from the uncoupled oscillation frequency f_0. It is further evident

from the form of (III.4) that y reaches extreme values $\pm \frac{1}{4} Q_1 k^2/(1-k^2)$ when $y-x$ is $\pm 1/(2Q_1)$. The full-line curves of Fig. III.-15 are plots of change in oscillation frequency of the circuit of Fig. III.-14 against detuning of the external circuit, for the various values of Q_1 noted and for a coupling coefficient of 2¼ per cent, such as is attainable in the usual magnetron. The dashed curve shows what would happen for a particular value of Q_1 with 22½ per cent coupling.

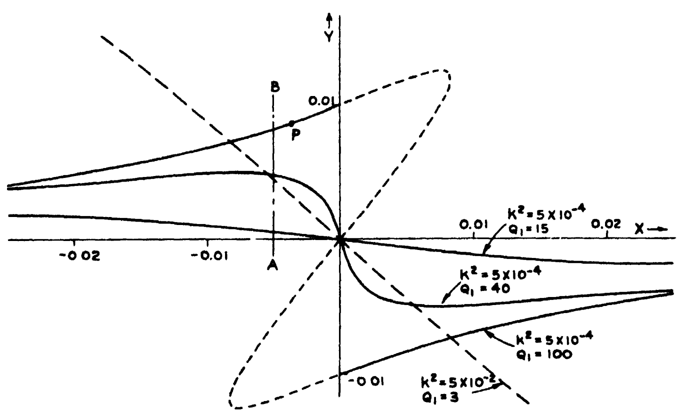

Fig. III.-15. Characteristics of frequency modulation using coupled circuit.

The dotted portion of the curve for $Q_1 = 100$ is actually plotted from the skewed cubic equation (III.4) but has obviously no physical meaning. In this region there are for each value of f_1 three frequencies f for which total susceptance vanishes, but of course the system can actually oscillate at any time at only one of these values of f. Oscillation will occur at that one of the three permitted frequencies for which the losses to be made up by energy drawn from the electron stream of the magnetron are least. This is that frequency satisfying equation (III.4) for which total susceptance G as given by equation (III.5) is smallest. Using values of x and y from the curve of Fig. III.-15 in the conductance equation, it is found that oscillations will actually take place only as indicated by the full-line portion of that curve. This behavior, with a break in oscillation frequency as f_1 is varied through f_0, is characteristic for sufficiently high Q of the coupled

circuit and is found experimentally. For lower Q_1, the figure shows that only a single frequency gives zero susceptance and no discontinuity in operation occurs. The value of Q_1 above which three possible values of y occur for certain values of x is $\sqrt{1-k^2}/k$.

An electron stream may be coupled to the external cavity so as to vary either its resonant frequency f_1 or its selectivity factor Q_1. Either effect provides a useful method for modulating the oscillation frequency of the magnetron. If Q_1 is varied, f will vary along a vertical line such as A-B of Fig. III.-15. If f_1 is varied, f will vary according to a full-line curve of the figure. It is obviously undesirable to vary f_1 through f_0 (x through zero) if Q_1 is high enough for a frequency jump to occur. If Q_1 is not quite high enough to produce a discontinuity, it is clear from the figure that there will be a small region near f_0 in which f varies rather rapidly with f_1. Operation in this region by varying f_1 is nevertheless undesirable because power loss in the modulating cavity is high, because it is difficult to insure linear modulation and because modulation sensitivity df/df_1 varies rapidly with Q_1, which is not a highly stable parameter. When modulating by variation of resonant frequency f_1 of the external circuit, it is therefore necessary to operate about such a point as P of the figure. The steepest ideally realizable characteristic for such operation, the limit approached in actual operation, comes at the zero value of x, with infinite Q_1. From equation (III.4), this corresponds to a value of $k/(2\sqrt{1-k^2})$ for y and to a limiting modulation sensitivity

$$(df/df_1)_{max} = (dy/dx)_{max} = ½; \qquad \text{(III.6)}$$

any actual modulation must occur with lower sensitivity than this. This is an important, but not necessarily prohibitive, limitation on frequency modulation of a cavity magnetron by an external singly-resonant "reactance tube."

Modulation by control of Q_1 has been used with moderate success.[20] For this purpose, the external cavity was loaded with a plane-electrode or "light house" diode and modulation was produced by varying a negative bias voltage applied to the diode. Q_1 was found to vary in normal modu-

lation between 40 and 150, with resulting frequency changes in good agreement with the predictions of equation (III.4). Due to long electron-transit time in the diode, bombardment by electrons accelerated by the r-f signal was more than adequate to maintain the diode cathode at emitting temperature, permitting the normal heater to be turned off. It was found by experiment that magnetron-output frequency varies non-linearly with either modulating voltage across or current through the diode, but in such a fashion that by shunting the diode with a suitable resistor a substantially linear variation of frequency with total current through diode and shunt may be attained over a limited but useful frequency range.

Amplitude modulation resulting from variation of Q_1 was found to be extremely slight, even with frequency shifts in excess of 5 megacycles, though the initial act of coupling on the modulating cavity markedly reduced the average power output of the magnetron. The load was coupled to the magnetron by a coupling loop separate from that used to couple to the modulator. Variations in reactive component of the load as seen by the magnetron cavity serve to change the tuning of that cavity and therefore its resonant frequency f_0. Reference to Fig. III.-15 shows that the modulation due to variation of Q_1 depends critically on the value of x at which operation takes place, that is on the small difference between uncoupled resonant frequencies f_0 of the magnetron and f_1 of the modulating cavity. It is therefore not surprising that the modulation characteristic was found to be affected strongly by variations in magnetron load. This is, in fact, a serious fault of the diode modulation system; another fault is the inherent non-linearity, which must be minimized by a carefully adjusted shunt resistor across the diode.

Limited investigation of the case of two resonant circuits coupled to the magnetron and both varied in resistance or reactance reveals useful possibilities. Under workable conditions, the change in frequency of oscillation still must be less than the change in coupled-circuit resonant frequency, but the frequency-change ratio may be somewhat greater than the limit of ½ found for a single circuit. It is, however, possible to produce a

usable working region, corresponding to the low-Q region near $x = 0$ of Fig. III.-15, in which oscillation frequency f varies sensibly linearly with simultaneous change of coupled-circuit resonant frequencies f_1 and f_2. This same region also has the property that, if modulation is produced by alteration of Q_1 and Q_2 rather than of f_1 and f_2, the modulation sensitivity changes only slightly for small changes of magnetron-cavity resonance f_0. A doubly resonant push-pull diode modulator may therefore be freed from the troublesome effects of variations in magnetron-load impedance.

It would be simplest to couple both load and modulating circuits to the magnetron by a single loop, but this seems only practicable if tight coupling can be obtained. The dashed curve of Fig. III.-15 shows that with sufficiently close coupling very modest values of Q_1 are adequate to permit strong control of oscillation frequency by the external resonant circuit. But coupling methods often used permit only weak coupling to the magnetron, and the solid curves of the figure show that for weak coupling a fairly high value of Q_1 is necessary to permit control. If the load is coupled directly to the modulating circuit in order to utilize a single coupling to the magnetron, the circuit Q is degraded by power dissipation in the useful load. The overall result with available coupling factors andunloaded-circuit Q values is that the effective value of Q_1 is too low for satisfactory modulation and the fraction of total generated power wasted in the modulator is too large.

Taken altogether, magnetron frequency modulation by variation of external coupled circuits is a practical and useful possibility which may well merit further development.

e. *Auxiliary Beam Tubes.* Magnetron oscillators may also be frequency modulated by varying the resonant frequency of their internal cavity structures. To do this mechanically at the highest modulation frequencies useful for f-m radar hardly appears attractive. There remains the possibility of direct control of resonant frequency of the magnetron anode cavities by auxiliary electron beams within the vacuum envelope of the magnetron itself. This involves the production of rather special tubes but has

proved a very useful method of frequency modulation.

If a beam of electrons is injected between the plates of a parallel-plate capacitor to which an alternating voltage of amplitude E is applied, as shown in Fig. III.-16, and a uniform magnetic field H is applied in a direction parallel to that of electron injection, the electrons will move in corkscrew fashion along the magnetic lines of force. The

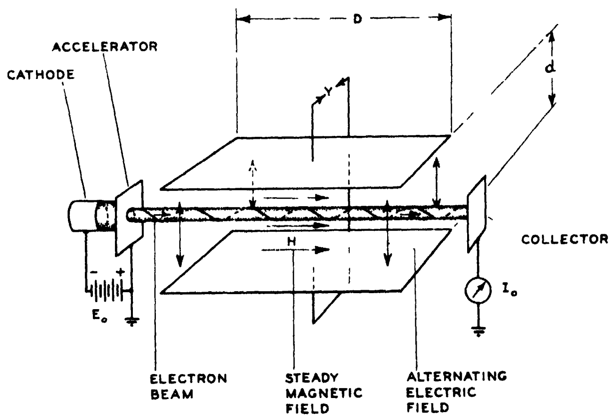

Fig. III.-16. Electronic impedance control.

electrons will have an oscillatory component of motion in the direction of capacitor-plate separation, representing a displacement current through the capacitor itself and a corresponding alternating current component in the external circuit connected to the capacitor.[21]

An electron injected into a uniform magnetic field of strength H with a velocity component perpendicular to that field will, in the absence of electric fields, travel along a helical path with axis parallel to the field. It will revolve around the axis of the helix with constant angular velocity

$$\omega_c = He/m, \qquad \text{(III.7)}$$

corresponding to the *cyclotron frequency* f_c, where e is electronic charge and m electronic mass. This frequency does not depend on velocity of the electron or radius of its helical path.

If now an alternating electrical field of the above frequency f_c is applied perpendicular to the magnetic field, an

electron injected parallel to the latter field will develop a periodic motion transverse to the magnetic field and in time phase with the electric field. It will remain in phase and be steadily accelerated, traveling in a steadily widening spiral and absorbing energy from the alternating source which supplies the field. The synchronous electron beam, though it may never strike the plates producing the electric field and so never cause an actual flow of electrons into a plate, will thus appear as a resistive impedance to the external driving circuit.

In the case of an electric field at a frequency f higher than the cyclotron frequency, an electron which begins moving in phase with the field when it enters the space between the exciting plates will initially pick up energy from the field and move in a widening spiral. At the same time, revolving at the cyclotron frequency f_c, it will begin to fall behind the field in phase and will lag more and more until, when it moves in phase quadrature with the field, energy exchange no longer occurs. Thereafter, the electron motion will have a component directly out of phase with the field and will be decelerated, moving then in a spiral of decreasing radius and falling still further behind the field in phase, while delivering energy to the external circuit. When the electron motion has just reached phase opposition to the electric field, the electron will have returned to the circuit just as much energy as it previously absorbed. If it passes out of the field at that moment, its presence will have resulted in no net energy transfer and so will not have placed a resistive load on the external circuit.

Throughout the above motion, the electron velocity will lag behind the field in phase by some amount, so that there will always be a lagging component of displacement current and the net effect of the electron on the external circuit will be that of an inductive reactance shunting the field-producing plates. In similar fashion, a field of frequency lower than the cyclotron frequency will be led in phase by the electron motion and the external circuit will in that case see the electron as a capacitive reactance.

By integrating the dynamic equations of electron motion in crossed magnetostatic and alternating electric fields,

for electrons entering the field region with velocity parallel to the magnetic field, the velocity component in the direction of the electric field may be determined. From this transverse velocity, with beam current I_o and electron velocity along the beam produced by a steady accelerating voltage E_o, the contribution of each length element of the beam to the displacement current between the plates may be determined.

These current contributions may be integrated over the length D of the interaction of beam and field to determine the total current to the plates due to the presence of the electron beam, and thence the shunt admittance Y seen by the external circuit. Calling T the time of transit of an electron over the distance D and θ the total lag $(\omega_c - \omega)T$ of electric-field phase behind electron motion during transit, and making some simplifications allowed by the fact that for the interesting conditions of operation $|f_c - f| \ll f$, the above procedure yields for the electronic admittance

$$Y_e = G_e + jB_e = \frac{I_o}{E_o} \cdot \frac{D^2}{4d^2} \left\{ \frac{1 - \cos\theta}{\theta^2} + j\,\frac{\theta - \sin\theta}{\theta^2} \right\}, \quad \text{(III.8)}$$

where d is plate separation.

If the capacitor threaded by the electron beam forms an element of a parallel circuit L_o, C_o, resonant with no beam at frequency f_o, the condition for resonance at a new frequency with the beam present will be that the sum of electronic and circuit susceptances shall vanish. Using B_e as given above, and considering the case for which $\Delta f = f - f_o$ is negligible compared to $f_c - f_o$, this condition leads to

$$\Delta f = \frac{I_o}{E_o C_o} \cdot \frac{D^2}{16 \pi d^2} \cdot \frac{\sin\theta - \theta}{\theta^2} \quad , \quad \text{(III.9)}$$

with shunt conductance

$$G_e = \frac{I_o}{E_o} \cdot \frac{D^2}{4d^2} \cdot \frac{1 - \cos\theta}{\theta^2} \quad . \quad \text{(III.10)}$$

Fig. III.-17 shows the dependence of Δf and G_e on phase difference θ accumulated in transit. Electron-field phase

difference θ is proportional to electron detuning $(f_c - f_0)/f_0$, the proportionality factor being the total electron transit angle at frequency f_0. In agreement with the qualitative earlier discussion, G_e is a maximum and Δf zero if $f_0 = f_c$ and no phase difference accumulates. Zero energy loss, with substantial Δf, occurs for one full cycle of phase lag.

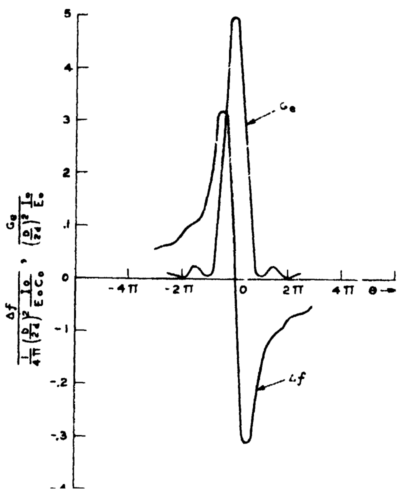

Fig. III.-17. Theoretical frequency shift and loading by helical electron beam.

Experiments at 4000 megacycles per second with an electron beam in one cavity of a vane-type multicavity resonator confirm the behavior suggested by equation (III.9). Within reasonable limits of error, theory and experiment agree in indicating a maximum frequency change of 0.3 megacycles per second per milliampere of an 80-volt electron beam. The maximum possible frequency shift, for a space-charge limited beam which just fails to strike the plates at the widest part of its motion, is found theoretically to be

$$\Delta f_{max} = {}^{2}\!/_{9}\, F(E_0/E)(C/C_0)f\,, \qquad \text{(III.11)}$$

where F is a factor between $^{7}\!/_{8}$ and 2 which depends on ratio of beam thickness to plate separation, C_0 is total effective capacitance, C is capacitance between plates, E_0 is beam-accelerating voltage and E is amplitude of alternating voltage.

Analysis of the behavior of an intensity-modulated

electron beam leads to much more complicated results[21] than those given above for a beam of constant intensity. Its properties depend not only on phase difference θ between electron and electric field during transit, but on modulation-frequency phase progress during transit also. The steady-state discussion given here still applies for modulation at sufficiently low frequencies. In the special case of $\theta = 2\pi$, it is found that the modulation frequency for a 4000-megacycle circuit may be raised to 300 megacycles before the frequency modulation obtainable with a given beam voltage and current drops to half its low-frequency value.

For the actual case of a distributed-constant resonant cavity, the analysis required involves solution of Maxwell's equations under appropriate boundary conditions by a method of successive approximations. Such an analysis has been carried out[21] for a rectangular-parallelopiped cavity with electron beam passing along the longitudinal center line of the cavity, with results broadly similar to but differing somewhat in detail from those found for the lumped circuit. For instance, maximum frequency shift occurs for phase difference in transit of 4 rather than π radians, while minimum loss occurs at 3π rather than 2π radians and minimum-loss operation permits 0.4 rather than 0.5 of maximum frequency shift.

The very important result of this work is that by approximately resonating the cyclotron frequency of an electron stream in a magnetic field with the frequency of an oscillatory circuit coupled to that stream, the reactive effect of the electrons on the circuit can be tremendously enhanced without introducing excessive losses. A further valuable feature is that the resulting frequency shift varies linearly with beam current, at least over a useful operating range.

On the basis of the foregoing principles and results, a special frequency-modulated magnetron[22] was developed, under the experimental type designation *A-127A*. Fig. III.-18 is an axial section showing the internal construction of the magnetron and in particular the grid-controlled sources of the modulating beams. Fig. III.-19 is a diagrammatic view along the axis of the strapped-vane

multicavity resonant anode of this magnetron, looking along the magnetic field and the modulating electron beams, which shows the relative size and position of the two auxiliary beams used.

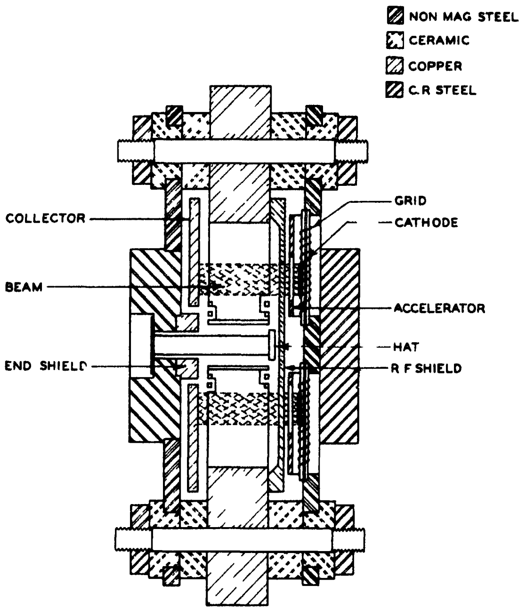

Fig. III.-18. Sectional view of structure of auxiliary-beam magnetron.

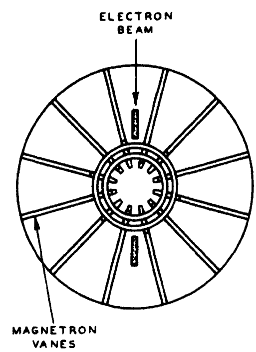

Fig. III.-19. Axial view of magnetron cavity and electron beams.

In the photograph of the magnetron structure with its metallic "bath tub" envelope removed, Fig. III.-20, the beam-forming elements are prominently visible, and between them the end "hat" of the magnetron cathode may be seen surrounded by the tips of the anode vanes. The radio-frequency shield necessary to protect the beam-cathode heaters against burn-out by induced r-f current is absent in this photograph. The load-coupling loop and its 50-ohm coaxial lead-out are quite normal and so not especially shown in these figures; the load is coupled to a cavity, at the bottom of the photograph, midway between those traversed by the modulating beams. All internal parts are mounted from and all leads brought out through the header plate, visible in the photo, to which the metal envelope will be welded. Massive copper rods supporting

the anode block serve to conduct heat to external cooling fins.

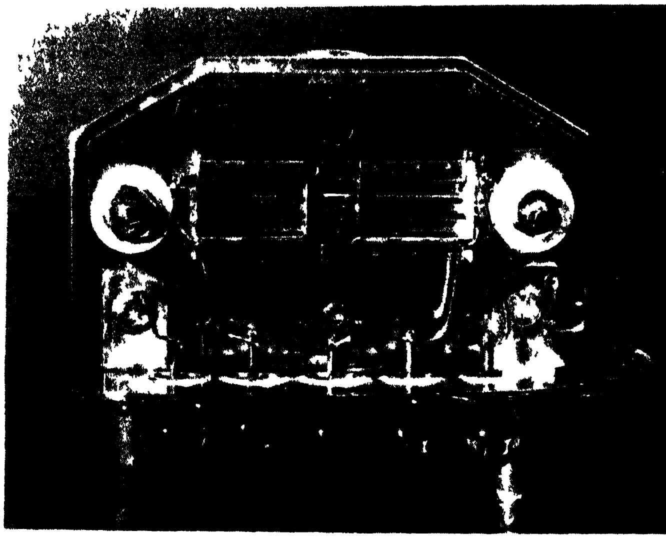

Fig. III.-20. Internal construction of 4000-megacycle auxiliary-beam magnetron.

Measurements on a number of *A-127A* magnetrons operating with 50-per cent efficiency and delivering 25 watts output into an approximately matched load while using 80-volt auxiliary beams are typified by the graphs of Fig. III.-21. Fig. III.-21(a) shows the effect of beam resonance on frequency modulation and power loss or resistive loading. These measured characteristics are seen to be qualitatively similar to the simplified theoretical characteristics of Fig. III.-17, to which they are in principle comparable, though differing in a number of details. These differences have not been fully explained, but are thought to result from effects of the very intense radio-frequency fields in the oscillating magnetron on beam-electron transit time.

Special test equipment[23, 24] was developed for obtaining dynamic frequency-shift characteristics of these tubes. In this equipment, beats of the frequency-modulated signal against a series of fixed-frequency reference signals were displayed as vertical pips on an oscilloscope trace, with horizontal deflection supplied by the modulating voltage. A voltmeter used in conjunction with an adjustable gate

displayed on the oscilloscope trace permitted correlation of instantaneous voltage on the modulating-beam grids of the magnetron with instantaneous output frequency. Fig. III.-21(b) shows the linearity of frequency characteristic obtained. A frequency swing of 3 megacycles was attainable with negligible amplitude modulation, with an 8-megacycle swing possible if 5-per-cent amplitude modulation may be tolerated. Use of modulating beams in more of

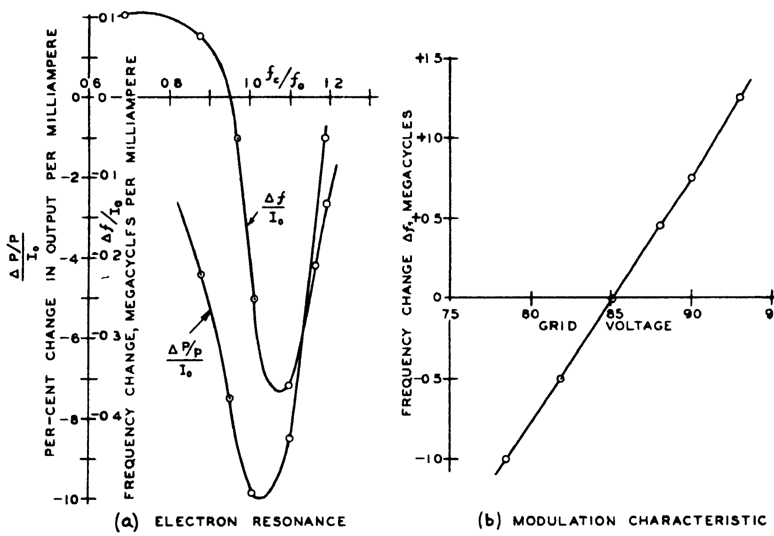

Fig. III.-21. Characteristics of auxiliary-beam control of magnetron.

the magnetron cavities would further increase the attainable frequency swing. These auxiliary-beam magnetron tubes are believed to be the first compact, efficient source capable of delivering significant amounts of linearly frequency-modulated power at super-high frequencies. Another auxiliary-beam magnetron,[25] producing much greater frequency-modulated power at rather lower frequency, has also been developed.

5. RECEIVING EQUIPMENT

a. *Balanced Detector.* In frequency-modulated radar as in other services, the simplest receiver is a detector fed directly with antenna output and in turn feeding a low-frequency amplifier. Because channel congestion has not

been serious in f-m radar operation, such a simple, unselective receiver has been adequate for some applications. If the detector is of the simplest type, however, radio-frequency signals reaching the receiver directly from the transmitter may produce overwhelmingly strong low-frequency interference in the detector output in case the transmitter is subject to some fortuitous amplitude modulation, as actual transmitters always are.

Balanced detectors have found use, because they permit a beat-note signal to be derived by mixing the receiving-antenna output with a local signal obtained from the transmitter, while at the same time they prevent any modulation-frequency signal from being produced by amplitude-modulated signals from either receiving antenna alone or transmitter

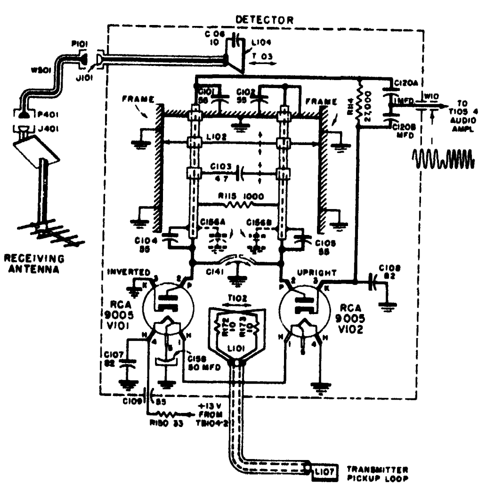

Fig. III.-22. Circuit diagram of balanced detector for 410-megacycle operation.

Fig. III.-23. Construction of 410-megacycle balanced detector.

alone. Fig. III.-22 is a circuit diagram of such a detector for 410-megacycle operation, and Fig. III.-23 is a photograph of the mechanical arrangement of the two acorn diodes and antenna-coupling elements. The antenna-coupling loop is inside the metal box at the center of the picture, while the local-signal coupling loop is in a similar box below the chassis deck. The resonant-line detector-input

circuit is adjacent to the chassis and between the two coupling loops; its ends may be seen connected to diode-socket terminals below the coaxial antenna line.

Referring to the circuit diagram, the parallel-resonant mode of the two input-line conductors is tuned by the adjustable shorting bar L-102 and excited by the antenna-coupling loop L-104, which together with its shielding box may be moved to equalize the coupling to the two line conductors. The push-pull mode of the twin line conductors is tuned by moving capacitive shunt C-103 and excited by the mixing-signal coupling loop L-101, which may also be moved laterally to balance the coupling to the two line conductors. The mixing-signal coupling loop, fed through a balanced shielded line from a pick-up loop coupled to the transmitter (see Fig. III.-8), is made highly symmetrical to avoid exciting the parallel mode of the detector-input line as a result of unbalanced current caused by stray capacitive coupling of the pickup loop to the transmitter.

The two diodes are connected in series with respect to their direct-current load R-114, and are so driven by the two input-line conductors that input from either transmitter or receiving antenna alone results in application of a balanced-to-ground, push-pull rectified signal across R-114. When both inputs are present at once, however, the beat-note output of the detector appears against ground at both ends of R-114 in parallel. The low-frequency output from the center-tap of the filter capacitor across the load resistor therefore represents beat note only and is free from modulation existing on either signal when received alone.

Balancing is accomplished by adjusting position of the two coupling loops and setting of the differential capacitor C-141. This is done with transmitter operating under frequency modulation and normally loaded, but with various impedances connected in place of the receiving antenna. A combination of settings is sought which will minimize audio output over as wide a range of receiver-input loading as possible. Use of the circuit-damping resistor R-115 makes the balance condition much less critical. At any one frequency, a high degree of balance against both detector inputs may be obtained, but when modulating over a band of

frequencies on optimum compromise is all that is possible. Even such a compromise is a matter of rather lengthy and critical adjustment. Changes of load seen by receiver input are likely to cause deterioration of balance. The balancing capacitor C-141 is intended to equalize effective diode capacitances, but the connections from it to the internal diode capacitances have appreciable inductive reactance. The capacitor setting for balance therefore varies with frequency; this is typical of the imperfections which prevent attainment of a uniformly high degree of balance over a wide frequency band.

When operating well, the balanced-detector system greatly reduces the disturbing effect of amplitude-modulated interfering signals and of amplitude modulation of the f-m radar transmitter. It is quite simple in construction, but critical adjustments must be made and maintained to secure satisfactory operation. Selectivity of the push-pull detector-input circuit coupled directly to the transmitter may convert the frequency modulation of the very strong directly coupled mixing signal to amplitude modulation. In case of imperfect balance, this modulation will appear directly as low-frequency detector output. Such converted modulation may be much stronger than the fortuitous amplitude modulation of the mixing signal which the balanced detector was introduced to suppress. A poorly adjusted and excessively selective balanced detector may therefore operate at least as badly as a good, non-selective single detector. This is the troublesome effect reduced by the use of push-pull damping resistor R-115.

In section 3 of Chapter II, the behavior of f-m radar was described in terms of amplitude variation of the resultant of direct and target-reflected radio-wave fields in the neighborhood of the radar transmitter. The significant variations of resultant field correspond to motion of a standing-wave pattern, caused either by changing transmitter frequency or by target motion. Actually, amplitude variation of the field is also likely to result from fortuitous amplitude modulation of the transmitter. Because the direct-signal component is usually many times stronger than the target-reflected component in the vicinity of the transmitter, spurious amplitude modulation of the resultant

field is likely to mask the much weaker modulation representing motion past the radar of the standing-wave pattern set up by the target. Special measures are required to overcome this unsatisfactory condition.

By carefully locating the receiving antenna so that the direct signal from the transmitter is minimized, as described in section 2b above, fortuitous amplitude modulation of the total received signal is very greatly reduced. By separately feeding a signal component directly from the transmitter to a balanced detector system, the effect of amplitude modulation of this component is also considerably reduced. These two artifices together serve to produce a detector output which much more nearly represents the standing-wave variations of the ideal case of Chapter II. than does the actual total field at most points near the real radar transmitter.

Balanced detectors have not been developed for 1500-megacycle operation, where it might be very difficult to maintain adequate balance. A special diode has been developed for such use,[26] however, incorporating provision for external control of anode-cathode spacing to permit accurate balancing of diode capacitances. At super-high frequencies, balanced detectors using crystal diodes and "Magic Tee" wave-guide circuits have been found useful.

b. *Side Band Superheterodyne*. The superheterodyne principle leads to a less critical if more complex receiver than the simple balanced detector, and particularly facilitates operation at the higher frequencies. This principle, however, requires rather special adaptation to meet the needs of f-m radar. Useful f-m radar information resides only in the small frequency differences between transmitted signal and returning echo, so that the wide frequency swing of the transmitted signal is no longer useful after the beat signal between it and the radar echo has been developed. It is therefore desirable that the local heterodyning signal follow the frequency sweep of the transmitter, so that the pass band of the intermediate-frequency amplifier need not accept the wide sweep of the returned signal but only the narrow spread between transmitted and received signals.

One method of making the heterodyne-signal frequency follow the transmitter frequency is to derive the former from the latter by a modulation process, using a filter to reject unwanted modulation components. This method may be described as a side-band superheterodyne. It has been used very successfully at both 515 and 1500 megacycles; the higher-frequency equipment in particular will be discussed here.

Fig. III.-24 is a block diagram of a side-band superheterodyne radar system. Transmission covers the band 1500 ± ½W megacycles per second, where W is width of frequency band swept in modulation. A portion of the

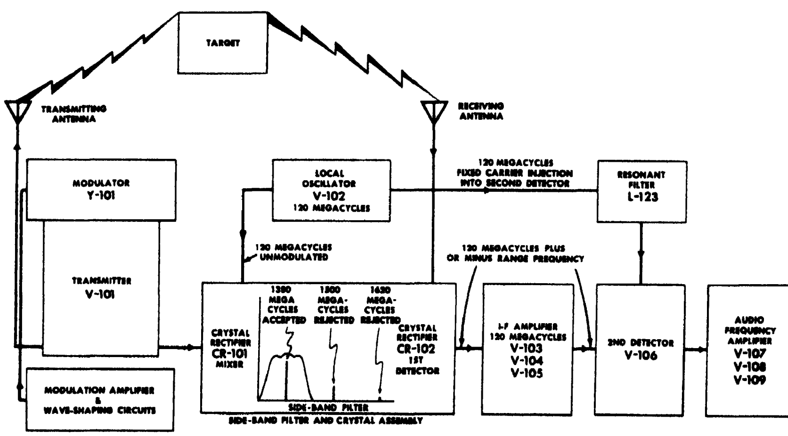

Fig. III.-24. Block diagram of side-band superheterodyne.

transmitter output is modulated in a crystal mixer by a 120-megacycle intermediate-frequency local oscillator. Mixer output feeds a single-side-band filter which strongly rejects both the carrier at 1500 ± ½W megacycles and the upper side band at 1620 ± ½W megacycles resulting from the modulation process, while freely passing the lower side band at 1380 ± ½W megacycles. The lower side frequency so passed of course tracks accurately the transmitter modulation and is suitable for use as a local heterodyning signal. This signal is mixed with the received radar-echo signal in a crystal first detector to produce a difference frequency of 120 megacycles ± $f_R + f_s$, where f_R and f_s are radar range frequency and Doppler speed frequency. The

difference-frequency heterodyned signal is amplified by a rather narrow-band intermediate-frequency amplifier and mixed in a diode second detector with 120-megacycle unmodulated signal from the local oscillator. Second-detector output is at the desired radar beat frequency $f_R \pm f_S$ and is fed to a low-frequency amplifier. In the 515-megacycle version the intermediate frequency is 30 megacycles per second, with the local i-f oscillator serving also as a first mixer.

The heart of this system is the side-band filter, which with the two crystal mixers forms the 1500-megacycle unit shown in Fig. III.-25. Input from the transmitter is applied, through the fitting H-104 at the right of the interior view, to a first circuit tuned to 1500 megacycles by the adjustable stub C-106. The crystal mixer serving to

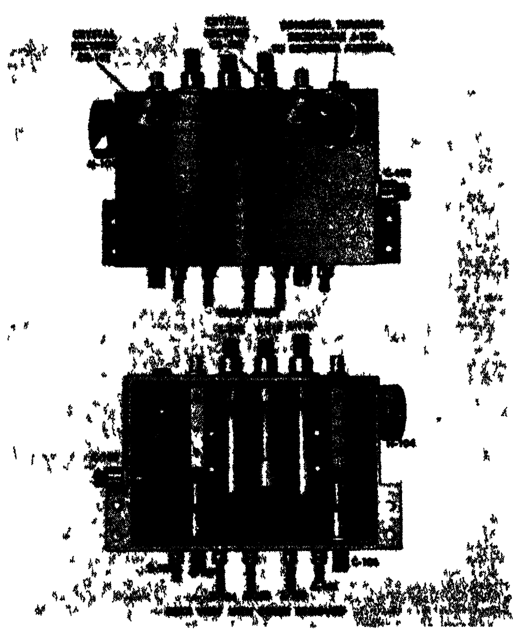

Fig. III.-25. Side-band filter and mixer unit for 1500-megacycle operation.

modulate this signal is connected through a mica blocking capacitor from the resonant rod of this circuit to the side wall of the filter housing and supplied with 120-megacycle modulating signal from the local oscillator. The second,

third, and fourth rods from the right are the coupled adjustably tuned circuits L-116, L-117 and L-118 of the 1380-megacycle band-pass filter proper, which has a pass band 24 megacycles wide to allow both for modulation and for frequency drift of the transmitter.

The first rod from the left is a circuit driven at a tapping point of suitable impedance by the coaxial line from the receiving antenna and tuned to 1500 megacycles by the adjustable stub C-108. Second from the left is a circuit which is tuned to 1500 megacycles by the stub C-107 and is coupled both to the 1500-megacycle receiver-input circuit and to the 1380-megacycle heterodyne-signal filter. The *1N21* crystal rectifier serving as first detector is connected through a mica insulating capacitor from this mixing circuit to the wall of the filter housing, and from it is taken the 120-megacycle intermediate-frequency output of the unit.

Coupling between the six tuned circuits is controlled by the adjustable rods Z-101 to Z-105 to achieve the desired band-pass and input-loading characteristics. Mixed inductive and capacitive couplings of opposing sense are present between adjacent circuits. Enlargement of the end of the resonant rod of filter circuit L-117 alters the proportions of capacitive and inductive coupling. This rod is so dimensioned that its couplings of the two types are just equal and opposite at 1500 megacycles, while the inductive component predominates at 1380 megacycles. The filter is thus given a strong rejection band at the transmitter frequency. Four holes seen in each wall of the housing are for shorting rods between front and rear wall, which prevent propagation across the filter by wave-guide action of third and higher harmonics present in the transmitter output, without affecting normal operation of the filter. The side-band filter in its 515-megacycle version uses three ordinary coil-and-condenser circuits and works into a balanced-diode first detector.

Low-frequency signals developed at the mixer by stray amplitude modulation of transmitter or local oscillator are stopped by the side-band filter, while similar signals developed by the first dector due to modulation of the heterodyning signal do not pass the intermediate-frequency

amplifier. Thus, the interference-rejecting properties of the balanced detector are obtained without critical balance conditions, and the design convenience of intermediate-frequency gain is made available as well. Side-band filter adjustments have proved satisfactorily stable. Strong rejection of the transmitted frequency by the side-band filter is essential; leakage of such signal is equivalent to cross coupling between antennas, and may cause much trouble if appreciably modulated in amplitude. A special crystal-size diode[27] was developed for use as a side-band-forming mixer in this system to avoid crystal burn-out, but the opportune appearance of high-burn-out crystals made unnecessary the use of these diodes.

If the second detector, in which 120-megacycle signal output from the i-f amplifier is mixed with direct signal from the 120-megacycle local oscillator, is balanced against the local-oscillator signal, low-frequency detector output due to stray modulation of the local oscillator may also be avoided. Such balance against a single-frequency signal at 120 megacycles is stable and easily produced; it was used successfully in the 30-megacycle i-f system of the 515-megacycle equipment. In general, the side-band superheterodyne system is more complex in construction but less critical and more stable in operation than the simple balanced-detector system.[28] For use against multiple targets, the local heterodyning signal at both first and second detectors must be much greater than any received signal. This is necessary in order to avoid beat notes, representing "ghost" targets, caused by inter-modulation of signals received from the various real targets.

c. *Signal Following Superheterodyne.* Still another special type of superheterodyne well adapted to f-m radar use is one in which the local oscillator is made to follow the frequency modulation of the transmitter by automatic frequency control. This system has been used successfully in experimental 4000-megacycle equipment. Like the side-band superheterodyne, it discards the wide sweep of radio-signal frequency after this sweep has served its purpose, and so permits use of a relatively narrow-band i-f amplifier.

Fig. III.-26 is a block diagram of such a signal-follow-

ing superheterodyne. Signals from the frequency-modulated transmitter and the controllable-frequency local oscillator are applied to a crystal mixer and the difference-frequency output of the mixer is amplified. The amplified difference signal is applied to a frequency discriminator, which produces a positive or negative control-signal output according to whether the difference-signal frequency is higher or lower than 30 megacycles. This control signal is applied to the reflector electrode of the local oscillator, a reflex klystron, and serves to maintain its frequency always substantially 30 megacycles lower than the modulated transmitter frequency.

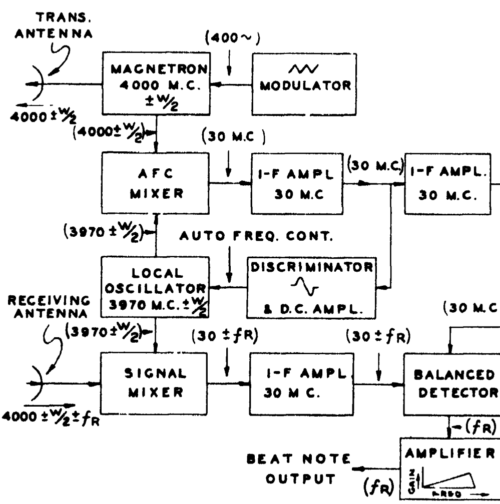

Fig. III.-26. Block diagram of signal-following superheterodyne.

Receiver signal and local-oscillator signal are applied to another mixer, the first detector of the receiver itself, which also develops output at a frequency of 30 megacycles. This intermediate-frequency signal is amplified and, along with the i-f signal from the control channel, is applied to still a third mixer, the second detector of the superheterodyne system. Second-detector output is at the desired radar range and Doppler speed frequencies.

Naturally, the automatic frequency control is not perfectly effective, since control signal is required and

can only occur if the frequency difference between transmitter and local oscillator departs somewhat from the intended 30 megacycles. This behavior is illustrated by graphs (a), (b), and (d) of Fig. III.-27, showing frequency variation with time for, respectively, transmitted and received signal, local oscillator, and transmitter-local-oscillator difference, which is the control-channel intermediate frequency. Similarly, graph (c) shows the received-signal-local-oscillator difference, the intermediate frequency in the signal channel; this is seen to depart cyclically from 30 megacycles as a result of

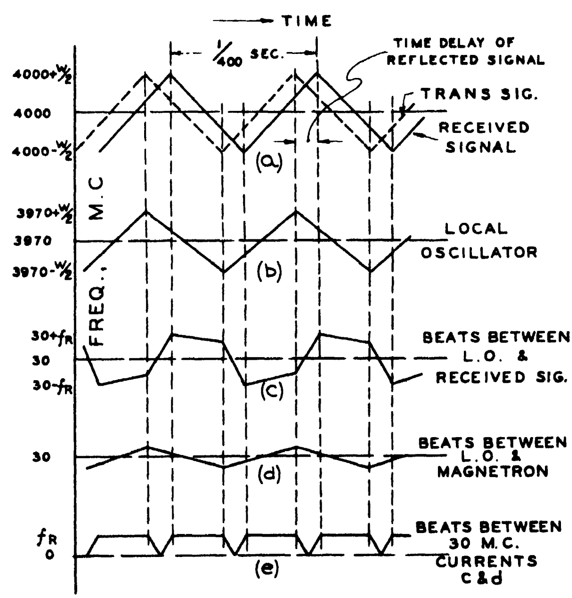

Fig. III.-27. Frequencies present in signal-following superheterodyne.

imperfect frequency control. In the second detector, however, signals (c) and (d) are combined and the signal-following imperfections of the local oscillator cancel from the final difference-frequency output shown by graph (e) of the figure.

Because of the intermediate-frequency modulation resulting from imperfect following, it is necessary that the two i-f amplifiers have the same phase-shift versus frequency characteristic. Failure to meet this condition will result

in spurious range beats when the amplifier outputs are combined. Non-linearity of the frequency-control characteristic of the local oscillator, however, does little harm because its effects are cancelled when the two intermediate frequencies are subtracted. As in the side-band superheterodyne, use of a balanced second detector eliminates certain types of interference or noise.

6. BEAT-FREQUENCY AMPLIFIERS

a. *Single-Target Systems.* The final element of the radio portion of an f-m radar system, as distinct from the indicator or other data-utilizing portion, is the amplifier for the (relatively) low-frequency beats between transmitted and target-echo signals. It is advantageous for this amplifier to have certain special characteristics, suited to the types of target and operation for which the system is intended.

If a single target is to be indicated at widely variable range, wide variations in received-signal strength are to be expected, along with variations in range-beat frequency. Signal-strength variations can in such cases be greatly reduced if the gain of the low-frequency amplifier is made to depend on frequency in the proper fashion. Range frequency varies linearly with range, while signal strength (amplitude) received by an altimeter varies inversely with altitude [see equation (II.34) or (II.35)]. By making amplifier gain increase linearly with increasing frequency, that is to say by giving the amplifier gain-frequency characteristic a slope of 6 decibels per octave (2:1 frequency change), the final signal-output level may be made independent of altitude above ground of constant reflecting characteristics. To continue this gain increase indefinitely, however, would increase the noise level without giving any compensating advantage. Amplifier gain should therefore be made to decrease with increasing frequency, as rapidly as is practicable, for all frequencies above that corresponding to the highest altitude to be indicated.

For a target of limited area, reflected-signal amplitude varies inversely as the square of the range [equation (II.33)], so a signal strength independent of range from such a target would require a beat-frequency gain increas-

ing with frequency at 12 decibels per octave. In practice, a compromise slope of about 9 db. per octave has been found most desirable for such targets. Again, a rapid gain cut above the maximum anticipated range frequency is desirable. If operation is required only at ranges exceeding a definite minimum value, a rapid decrease in gain is desirable for frequencies below that corresponding to minimum range. This is especially important because of the strong signals at very low frequencies that can be produced by microphonics, stray amplitude modulation of the transmitter, modulation of feed-through signal, or unwanted altitude signal.

To ensure proper operation of all vacuum tubes in the presence of strong signals, it is advisable to provide for automatic gain control actuated by the level of the final output signal. Where the desired rising gain-frequency characteristic is obtained by selective feed back, the further advantage of having the automatic gain control vary the shape as well as the level of the gain characteristic may be obtained. This results in selectively reducing gain at the highest frequencies, so reducing the noise bandwidth of the system and improving signal-to-noise ratio for strong signals. A further advantage of this automatic response control is that, in the presence of a strong nearby target, it discriminates against interference from distant targets which may happen to lie in the line of transmission. Response control to discriminate against distant targets and high-frequency noise may alternatively be based on range-beat frequency rather than signal strength, high-frequency gain being progressively reduced as target range decreases.

Gain-frequency characteristics of a typical beat-frequency amplifier for altimeter operation are shown by Fig. III.-28 for three values of gain-control bias. Strong reduction of gain at very low frequencies is obtained from properties of the input-coupling transformer fed by the detector, as well as by use of low-valued screen-grid bypass capacitors and inter-stage coupling capacitors. High-frequency gain cut is also obtained by use of input-transformer properties, as well as by shunt capacitance to ground in plate and grid circuits. Sloping response

in the operating region results from the application of degenerative feed back, through a low-pass circuit, to one amplifier stage, and reduction of gain by increased bias on this stage alone serves to change the shape of the overall characteristic in the way shown.

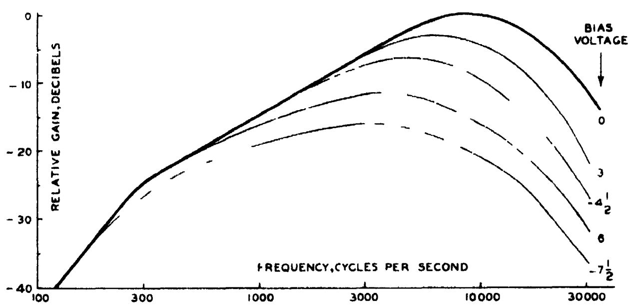

Fig. III.-28. Typical response characteristics of altimeter beat-frequency amplifier.

In another type of single-target operation, a control output is to be produced when a fairly definite range frequency occurs. This calls for a different gain-frequency characteristic. Useful signal is required only for ranges near that for control actuation, or sufficiently greater to permit circuit transients caused by initial signal rise to die out before such actuation occurs. The required characteristic is of a more or less flat-topped band-pass type, with the steepest practicable decrease in gain at frequencies above and more especially below the pass band. The pass-band width must be sufficient to accommodate anticipated shifts due to speed variation, and its limits should not depend upon signal level.

b. *Multiple Targets.* The properties of a beat-frequency amplifier for multiple-target indication will vary widely in accordance with the particular method of operation used, but will always remain those of a wave analyzer of some sort. If simultaneous analysis on all frequencies is required, many parallel-input, sharply tuned fixed selective channels will be needed, each one capable of signalling the presence or absence of a target within predetermined range limits. Each selective channel corresponds to a fixed range gate in pulse-radar technique. If

sequential analysis is permissible, then a single sharply selective amplifier must be made to scan the range-frequency band. This may for example be done by variable heterodyne means as in most commercial wave analyzers. Alternatively, a fixed-tuned amplifier may be used and the spectrum of radar range-beat frequencies may be made to scan across the narrow amplifier pass band, by variation of either the repetition frequency or the frequency swing of the radar modulation.

7. NOTATION AND REFERENCES

a. *Notation*. The algebraic notation listed alphabetically below has been used in this chapter.

B	Circuit susceptance.
B_e	Circuit susceptance resulting from motion of free electrons.
C	Circuit capacitance, usually with subscript to identify particular capacitor.
d	Separation of electron-deflecting plates.
D	Distance traveled by electrons in alternating electric field.
e	Charge of electron.
E	Amplitude of electron-deflecting voltage; also, supply voltage for modulation generator.
E_o	Electron-beam accelerating voltage.
f	Frequency.
f_o	Natural oscillation frequency of magnetron; also, resonant frequency of electron-beam deflecting circuit without electron beam.
f_1, f_2	Frequency characterising effect of resonant circuit coupled to magnetron (near resonant frequency of coupled circuit).
f_c	Cyclotron frequency of electron revolution in magnetic field.
f_m	Frequency of modulation.
f_R, f_S	Radar beat frequencies due to target range and speed respectively.
Δf	Change in resonant frequency produced by electron beam.
G	Circuit conductance.
G_o	Effective conductance of magnetron and coupled load.
G_e	Circuit conductance resulting from motion of free electrons.
H	Strength of magnetic field.
I_o	Electron beam current.

k	Coefficient of circuit coupling.
L	Circuit self inductance, usually with identifying subscript.
m	Mass of electron.
M	Mutual inductance.
Q	Selectivity factor of circuit, used with identifying subscript.
r	Circuit resistance, usually with identifying subscript.
T	Time of electron transit through field.
W	Width of band swept in frequency modulation.
x	Fractional detuning of circuit coupled to magnetron.
y	Fractional change in oscillation frequency produced by coupled circuit.
Y	Circuit admittance.
θ	Phase difference between electron motion and field alternation accumulated in transit of electron through field.
ω	$2\pi f$.

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CHAPTER IV.

APPARATUS FOR UTILIZATION OF F-M RADAR DATA

1. General Requirements

Output from the radio portion of frequency-modulated radar equipment is in the form of beat-note signals at frequencies determined by target range and relative speed of radar and target, as explained in Chapter II. This is complete data but is not in directly useful form. Such data must be converted to currents, voltages, shaft positions or relay operations for single targets, or into a frequency spectrum display for multiple targets, in order to be directly useful. The following discussion of means for data conversion and utilization will relate mostly to single-target operation, which has been investigated much more extensively than has the case of multiple targets.

2. Averaging Cycle-Rate Counters

a. *Basic Principle*. The principle of the data-converting element that has seen most use is very simple. This device serves to produce an output current of average value proportional to the frequency of the input and may be represented by the simple circuit of Fig. IV.-1. Switch

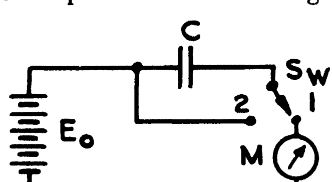


Fig. IV.-1. Basic cycle-rate counter.

Sw is driven synchronously by the signal to be measured, making contact in each position once per signal cycle.

When contact is made in position 1, capacitor *C* charges rapidly through meter *M* to voltage E_0 supplied by the

battery, receiving a charge

$$q = CE_0 \qquad \text{(IV. 1)}$$

When contact is made in position 2, this charge is discharged in a surge or pulse of current. This sequence of operations is repeated for each cycle of the frequency to be measured, or f times per second, so that a total charge fq passes through the meter per second. The average current through the meter is therefore

$$i = fCE_0 . \qquad \text{(IV. 2)}$$

A 100-micromicrofarad capacitor charged to 100 volts 100 times per second, for example, passes an average current of 1 microampere.

Since the meter current consists of a sequence of pulses, one per cycle of switch operation, and has an average value proportional to the time rate of recurrence of these cycles, the circuit of Fig. IV.-1 may be described as an "averaging cycle-rate counter." This circuit will hereafter be called for brevity simply a *counter*. It should not be confused with other devices, of the nature of electrical ratchets, to which the name counter is also applied. The averaging cycle-rate counter is an absolute frequency-measuring device, with accuracy depending only on the current-indicating accuracy of the meter and on the accuracy of the circuit element C and supply voltage E_0. So long as the capacitor has time to charge and discharge fully at each operation, duration of closing of the switch is immaterial.

Mechanical switching is of course impracticable for the measurement of any reasonably high frequency. An electronic analogue of the circuit of Fig. IV.-1 is therefore used, as shown by Fig. IV.-2. A strong signal at the frequency f to be measured is applied to the grid of the switching tube or *limiter* V_1. Capacitor C is either charged through resistor r_0, diode V_2 and meter M, or is discharged through limiter V_1 and diode V_3, as V_1 is alternately made non-conducting or highly conducting by the signal applied to its grid. The diodes provide separate channels for charging and discharging currents, so that either current alone may be passed through the measuring

instrument. Resistor r_o slows down the charging of C but, so long as V_1 always remains non-conducting for a sufficient interval to permit C to become fully charged, the value of r_o does not affect the counter output.

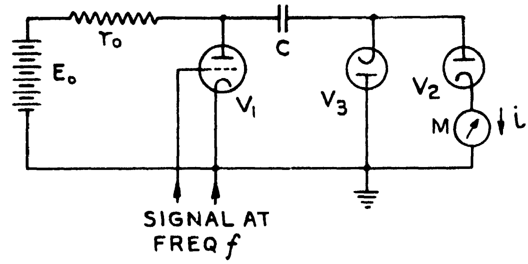


Fig. IV.-2. Electronic cycle-rate counter.

When one of these counters is fed a signal whose frequency alternates between two values, as does the beat-note output of a radar with symmetrical-sawtooth frequency modulation when operating against a moving target, the counter-output current will be a measure of the average of the two input frequencies. For target range and speed so related to the radar characteristics that range frequency exceeds speed frequency, the counter will measure range. For cases in which the speed frequency is the greater, the counter will measure speed. For more complicated beat-frequency variations, such as result from non-linear frequency modulation of a radar signal, the counter will still measure the time average of its input frequency.

Production of a current proportional to frequency by repeated charging of a capacitor is a very old and well known procedure. Circuits similar to that of Fig. IV.-2 are the basis for the wide range, direct-indicating frequency meters now marketed by many instrument manufacturers. One point of technique is important in their use: no appreciable leakage in either diode or in the counter capacitor is permissible.

b. *Non-Linear Form.* Currents of a few microamperes are not suitable for actuating rugged data-utilizing or indicating elements. Passage of the output current of an averaging cycle-rate counter through a large resistor

in order to develop a significant voltage, however, modifies the action of the counter. Fig. IV.-3 shows a counter developing a voltage output e across a load resistor r.

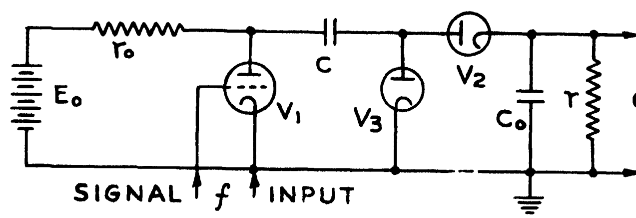


Fig. IV.-3. Non-linear counter.

Capacitor C_0 provides integrating action to average the current pulses from the counter so that a smoothed voltage will be developed; this replaces the smoothing effect provided, in the simpler current-indicating circuit of Fig. IV.-2, by mechanical inertia of the moving parts of the meter M.

Because of the presence of output voltage e, the net voltage available to charge capacitor C is only $E_0 - e$, so that

$$i = e/r = fC(E_0 - e). \tag{IV.3}$$

Solving for e, this gives

$$e = E_0 f/\left[f - 1/rC\right]. \tag{IV.4}$$

This voltage output varies in non-linear fashion with frequency, as shown by the graph of Fig. IV.-4.

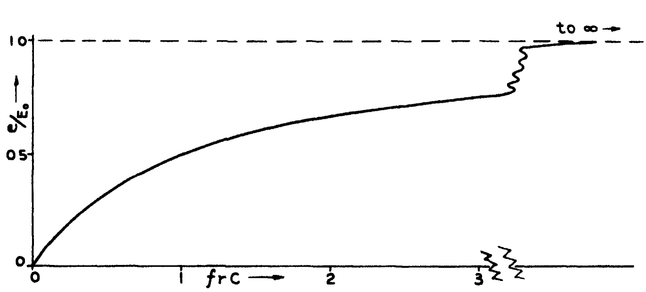


Fig. IV.-4. Output characteristic of non-linear counter.

The non-linear counter, with its high sensitivity at

low frequencies and saturation characteristic at high frequencies, may be quite useful for special purposes. In general, however, a linear characteristic is preferable.

c. *Linear Form.* A counter developing significant voltage across an output resistor may nevertheless be made to have an accurately linear relation between input frequency and output voltage. This is done by returning the discharge diode to a point maintained by some external means at the counter-output voltage, rather than to a fixed zero-voltage point. A cathode follower provides a simple and suitable means of maintaining a low-impedance return point at substantially counter-output voltage. Fig. IV.-5 shows the circuit of the linear counter so obtained, which

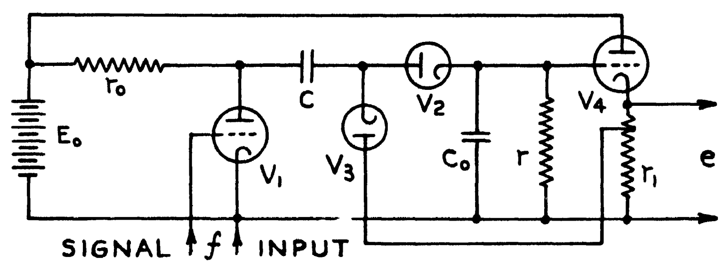


Fig. IV.-5. Linear counter with positive output.

has the further advantage of providing a useful voltage output at low impedance. While limiter V_1 is non-conducting, capacitor C becomes charged to voltage $E_0 - e$. While V_1 is heavily conducting, C becomes charged to voltage $0 - e$. The net change in charge per cycle is therefore CE_0, which is independent of output voltage e, and

$$e = frCE_0 . \qquad \text{(IV.5)}$$

In normal operation, the cathode of follower tube V_4 will run at a more positive voltage than the grid, in order to provide V_4 with suitable bias. Discharge diode V_3 must therefore be returned to a tap on cathode resistor r_1 in order to see a voltage equal to the output voltage e. Grid-cathode voltage on V_4 must vary to cause the cathode voltage to follow variations in grid voltage e; this imperfection in following action results in slightly imperfect counter linearity and may be serious in highly critical applications.

Referring to the simple circuit Fig. IV.-2, diodes V_2 and V_3 may be seen to be connected in series in a closed loop, with aiding polarity. Because of the internal voltage drop caused by finite electron-emission velocities and interelectrode contact potentials in each diode, a current would circulate in this loop even if the switch tube should remain permanently non-conducting. When switching, this current will prevent proper separation of capacitor charging and discharging circuits; it will therefore vitiate readings of meter M, and must be eliminated. This may be accomplished, in the circuit in question, by returning the plate of V_2 to a suitable fixed negative voltage rather than to ground.

With a cathode follower acting to maintain cathode of diode V_2 and anode of diode V_3 at the same voltage in the linear counter of Fig. IV.-5, a similar difficulty from diode-loop currents would occur. One remedy is to connect the counter-return lead to cathode resistor r_1 at a point having a voltage lower than grid voltage e by an amount exceeding the total internal potential of both diodes. Further departure from ideal operation occurs because switch tube V_1 never becomes perfectly conducting. V_1 therefore has some small residual voltage drop E_1 across it even when most conductive.

Taking the residual limiter drop into account, as well as the voltage difference e_d of cathode-follower grid above diode-return tap and the internal voltage e_i of each diode, the actual voltage to which C charges is $E_0 - (e - e_i)$ and its discharged voltage is $E_1 - (e - e_d + e_i)$. The net capacitor voltage change is

$$E'_0 = E_0 - E_1 - e_d + 2e_i, \qquad \text{(IV.6)}$$

which is again independent of e (e_d does actually depend somewhat on e as a result of cathode-follower action, however). It is E'_0 rather than E_0 which should really appear in equation (IV.5), but except for this change that equation remains valid for real, imperfect vacuum tubes.

In f-m radar use, for the case of range beat-note frequency greater than speed beat-note frequency, the average frequency throughout the modulation cycle is the range frequency f_R, and for such use this is the frequency

that should appear in (IV.5). The linear counter fed with such a signal may be called a *range counter* and, from (IV.5) using E_0' as given by (IV.6), will exhibit a *range counter sensitivity* h_R of

$$h_R = rC_R E_0' \qquad \text{(IV.7)}$$

volts per cycle per second. This sensitivity may be adjusted by varying any one of its three factors; each type of control has found practical use. For the case of speed frequency exceeding range frequency, on the other hand, the counter output would represent speed rather than range.

The linear counter of Fig. IV.-5 is a very useful and flexible device. Controlled non-linearity may be introduced by tapping the plate of discharging diode V_3 further down on the cathode resistor r_1 and making use of the consequent increased variation of e_d with e. The circuit as shown of course produces a positive voltage output. Negative incremental output due to operation of the counter may be produced instead by feeding load r and cathode-follower grid from the plate of V_3 rather than from the cathode of V_2 as shown. In this case, the cathode of charging diode V_2 will be returned directly to the cathode of V_4, so that the grid-cathode voltage of the latter will be applied with proper polarity to oppose the internal potentials of the now-reversed diodes. Equation (IV.6) for effective swing E_0' still holds for the negative-output counter, if e_d now represents voltage difference of follower cathode above follower grid. Counter load r will not be returned to ground but to a fixed positive voltage producing cathode-follower operation in a suitable plate-current range.

d. *Null Form*. Very accurate frequency measurement may be realized by using a null or slide-back counter circuit such as is shown by Fig. IV.-6. In this arrangement, a negative-output counter has its load resistor returned to a variable tap on a voltage divider or potentiometer r_2, fed from the same voltage source which supplies the plate of the limiter V_1. For any particular input frequency f, output voltage e may be made zero by adjusting the potentiometer to apply an appropriate fraction x of the supply voltage E_0 to the load resistor r. When e is zero and counter-output current is i,

$$i = fCE_o = xE_o / r \tag{IV.8}$$

and cancelling E_o,

$$x = frC. \tag{IV.9}$$

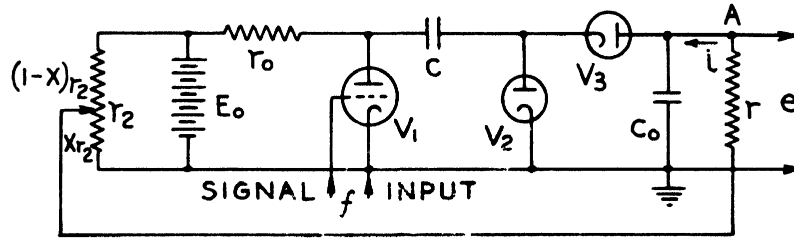


Fig. IV.-6. Null-type counter.

Frequency measurement with a null-type counter is seen not to depend on supply voltage E_o, a considerable practical advantage. Actually, the charging-diode cathode must be returned not to ground but to a positive voltage e_d sufficient to overcome the combined internal diode potentials and prevent parasitic current flow in the absence of limiter switching. The voltage applied to the potentiometer for proper compensation of supply-voltage variation should therefore not be just E_o but rather the actual voltage swing impressed on C, which corresponds to the E_o' of equation (IV.6).

So long as the current E_o'/r_2 through the potentiometer is much greater than the counter current i, the division ratio x of the potentiometer is independent of i and dependent only on potentiometer setting. Under this condition, accuracy of null frequency measurement depends only on accuracy of voltage-divider calibration and of the fixed circuit elements C and r, as well as on the sensitivity of the null indicator used when adjusting net output voltage to zero. Except in the matter of sensitivity, impedance of the null indicator is not important.

Of course, adjustment of the tap on r_2 to reduce e to zero may be made automatically and continuously by a servo mechanism. Load resistor r might alternatively be returned directly to supply voltage E_o and balance obtained by varying r. This would give a device with a reciprocal or hyperbolic rather than a linear frequency scale. Actually, with small vacuum tubes the

value of r must be of megohm order and the lack of any accurate and reliable variable high resistor renders such a modification impracticable at present.

For zero output voltage e, divider-ratio setting x varies strictly linearly with counter-input frequency f. For a constant divider setting, however, output e varies in non-linear fashion with input frequency. Taking account of any finite output e gives

$$fC(E_o' + e) = (xE_o' - e)/r \tag{IV.10}$$

instead of (IV.8). Equation (IV.10) may be solved for e to give

$$e = E_o' \left[f - {}^x/_{rC}\right] \left[f - {}^1/_{rC}\right] , \tag{IV.11}$$

which is rather similar to the output variation (IV.4) of the simple non-linear counter.

In null operation, this non-linearity represents only a dependence on frequency of null sensitivity to variation of either frequency or voltage-divider ratio; it can cause considerable inconvenience but does not impair accuracy except by decrease of indicator sensitivity. Even the addition of a linearizing cathode follower fails to prevent entirely this sensitivity change when operation over a wide frequency range is required. Sensitivity variation by no means prevents the null type of counter from being a very useful and generally satisfactory device.

e. *Effect of Smoothing*. In the case of steady input frequency, the smoothing capacitor C_o has no effect on counter output except to reduce voltage ripple at the frequency being measured. An important field of use of f-m radar, however, involves ranges decreasing uniformly with time and correspondingly requires determination of a frequency decreasing uniformly with time. For this case, the effect of C_o is more serious; it may most conveniently be investigated for the linear counter with cathode follower.

Counter-output current i, given by equation (IV.2) above and made independent of output voltage e by the cathode follower, flows into load resistor r and smoothing capacitor C_o in parallel (see Fig. IV.-5). Resistor current is

proportional to voltage and capacitor current to rate of change of voltage across the parallel combination, so that total current

$$i = \frac{e}{r} + C_o \frac{de}{dt} = fCE'_o \, . \qquad \text{(IV.12)}$$

Let f change with time at a uniform (negative) rate $\dot{f}$, reaching a reference value f_o at time t_o so that

$$f = f_o + \dot{f}(t - t_o) \, . \qquad \text{(IV.13)}$$

From (IV.12) and (IV.13),

$$\frac{de}{dt} + \frac{e}{rC_o} = \epsilon^{-t/rC_o} \frac{d}{dt}\left[e\epsilon^{t/rC_o}\right] = E'_o(f_o - \dot{f}t_o)\frac{C}{C_o} + E'_o\dot{f}t\frac{C}{C_o} \, . \qquad \text{(IV.14)}$$

This differential equation has the solution

$$e = f_o rCE'_o - \dot{f}r^2CC_oE'_o + \dot{f}(t - t_o)rCE'_o + K\epsilon^{-t/rC_o} , \qquad \text{(IV.15)}$$

where K is a constant of integration.

Let counting start at the zero of the time scale, which is so chosen that $t >> rC_o$ for the interval of interest near t_o. The transient exponential term of equation (IV.15) may then be neglected. For a steady input frequency f_o, counter output e would have the value $f_o rCE'_o$. From equation (IV.15), it is clear that in the case of uniformly decreasing frequency this value of output voltage will only be reached at time

$$t = t_o + rC_o , \qquad \text{(IV.16)}$$

although the frequency f_o was reached earlier at time t_o. That is, the presence of the smoothing capacitor causes the counter-output voltage to follow a linear time variation of input frequency with a constant time lag equal to the time constant rC_o of the load circuit, once the starting transient has died out.

f. *Limiter Circuits*. Limiter or switching tube V_1 has been shown in the preceding figures as a triode for simplicity, but such a tube is not actually satisfactory for this purpose. Even though electron plate current may be cut off by negative signal on the grid, further changes in grid voltage can still affect plate voltage of a triode slightly by direct capacitive coupling within the tube.

At the other end of the signal swing, the minimum plate voltage reaching the triode through feed resistor r_0 depends on the maximum voltage reached by the grid, so that plate-current limiting occurs only as a result of limiting the positive grid-voltage swing. By making the resistance of the grid-driving signal source very high, the positive swing of the grid voltage can indeed be limited by the onset of grid current; however, this still does not provide a highly definite minimum plate voltage to control the end point of the discharge of counter capacitor C.

In the case of pentode or beam-tetrode tubes, a much more favorable condition exists. Negative grid-voltage excursions beyond the point of plate-current cut off produce no appreciable effect on plate voltage, so that the counter capacitor may be charged just to the full supply voltage. Given a sufficiently large resistance r_0, substantial coincidence among the lower portions of the plate-current versus plate-voltage characteristics of such tubes for various grid voltages indicates that the minimum plate voltage reached is independent of the maximum grid voltage. This insures a definite voltage level to which the counter capacitor must discharge.

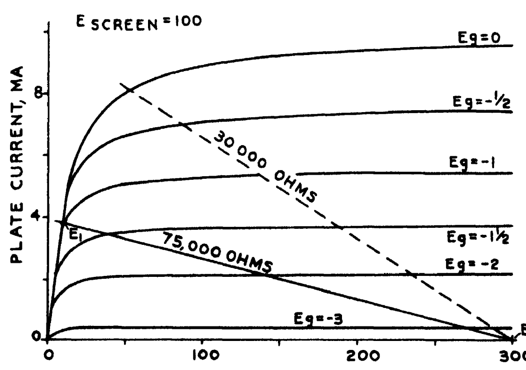


Fig. IV.-7. Pentode characteristics (6SH7 tube).

Fig. IV.-7 shows a typical family of pentode plate characteristics, together with the load line for a suitable feed resistor. The plate voltage occurring for any particular grid voltage is that at which load line and appropriate plate characteristic intersect; it may be seen to reach a

minimum E_1 which is independent of grid voltage for all grid voltages above -1. A lower feed resistance such as is represented by the dashed load line of the figure fails to produce such a definite result, plate voltage continuing to depend on grid voltage right up to the grid-current point.

Varying screen voltage alters the plate characteristics in much the same way as does varying grid voltage, so that minimum plate voltage is also reasonably independent of screen voltage. In fact, the pentode plate circuit at low voltage behaves very much like a simple resistor of definite and rather low value. With a pentode limiter driven by a large signal on the control grid, an accurately flat-topped plate-voltage wave is therefore produced. Plate voltage swings between the definite limit E_o set by the voltage supply and the definite limit E_1 set jointly by supply voltage, plate-feed resistance and tube characteristic. Two pentode-limiter stages in cascade give a still more accurate result.

To provide sufficient time at minimum plate voltage for complete discharge of the counter capacitor on each signal cycle, the voltage of a sinusoidal input signal must swing much more positive than the grid-current point of the limiter. This requires a high-impedance driving source in order to avoid excessive peak grid current. A large negative over-swing of grid voltage is also necessary to insure complete charging of the counter capacitor. In case capacitive coupling to the limiter grid is used, the unsymmetrical flow of grid current will make limiter bias depend on signal amplitude. This variation of bias is inconvenient and may affect unduly the exact lower limit of plate voltage. Bias change may be minimized by the input connection of Fig. IV.-8(a), with r_p much less than r_s. It may be practically eliminated by the circuit of Fig. IV.-8(b), in which the diode plate is returned to a fixed voltage lower than the plate-current cut-off voltage of the limiter grid, so that while the useful limiter-grid swing is not affected the diode current will balance the effect of the grid current.

Where accurate control of the limits of output-voltage swing is required, with output swing accurately proportional to supply voltage, the double cathode-follower

limiter of Fig. IV.-9 is useful. It also provides very low output impedance to insure full charging and discharging of counter capacitor C at high operating frequencies.

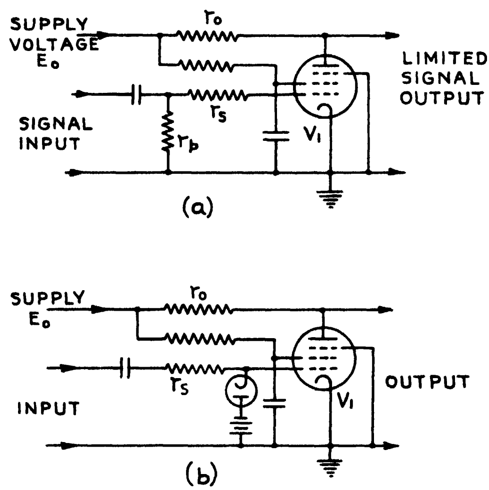


Fig. IV.-8. Limiter input circuits.

When limiter V_1 is cut off, output voltage is maintained at a predetermined fraction of supply voltage by V_a acting as a cathode follower; V_b is then cut off. When limiter V_1 is

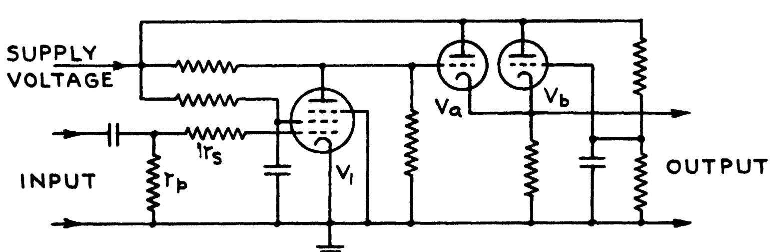


Fig. IV.-9. Precision limiter.

at minimum plate voltage, output voltage is maintained at a separately predetermined fraction of supply voltage by V_b acting as a cathode follower; V_a is then cut off. This type of limiter permits very accurate compensation of a null-type counter against supply-voltage variations.

In certain cases, noise or interfering signal of much higher frequency than the desired signal may be present in

disturbing amount. Action of a single limiter will equalize the amplitude of noise and signal and serious counting error may result because of the high frequency of the noise. By integrating the limiter output in a resistance-capacitance circuit, the relative amplitude of the noise may be greatly reduced. Applying the integrated first-limiter output to a second limiter, a square-wave output with the noise practically eliminated may be produced. Fig. IV.-10 shows wave forms

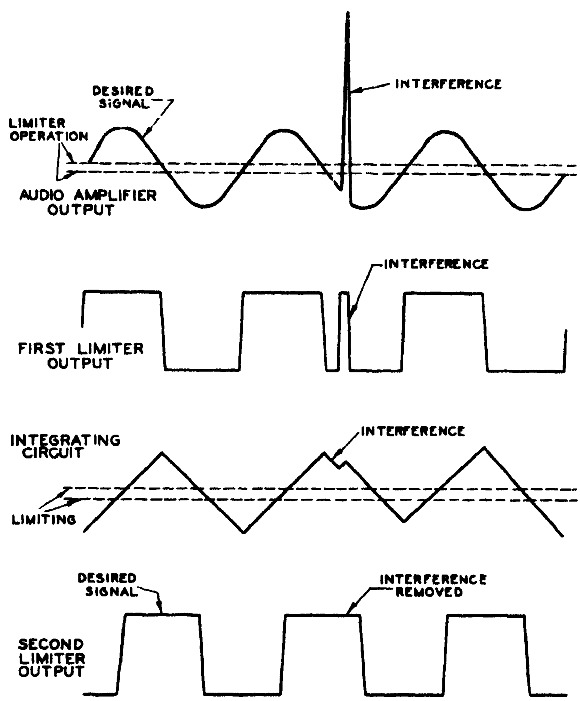


Fig. IV.-10. Operation of integrating double limiter.

present at various points in an integrating double limiter. This type of circuit is very useful in some cases but may be harmful in others. In the presence of strong interference or noise of much lower frequency than the desired signal, such as may result for example from a poorly balanced detector, the process described above will act to suppress the desired signal and emphasize the noise.

g. Fixed Error. Averaging cycle-rate counters indicate the average frequency with which the limiter-input voltage passes in one direction through the range between cut off and grid current. With large input signals and strong limiting, this must occur an integral number of times in each cycle of the modulation of an f-m radar, so that the momentary counting rate is necessarily an integral multiple of the modulation frequency. When range is strictly constant, this fact gives rise to systematic errors of indication,[1] already briefly mentioned in section 4k of Chapter II., which must now be studied more closely.

Consider a radar operating at average radio frequency F_o and modulated to sweep a total frequency band of width W. The number N_o of standing waves between radar and target at range R for frequency F_o will be

$$N_o = 2R/\lambda_o = (2/c)F_oR. \qquad \text{(IV. 17)}$$

The change ΔN in number of standing waves produced by the modulation sweep of width W will be

$$\Delta N = (2/c)(F_o + \tfrac{1}{2}W)R - (2/c)(F_o - \tfrac{1}{2}W)R = (2/c)WR. \qquad \text{(IV. 18)}$$

That is,

$$\Delta N = 2R/\lambda_w = N_w, \qquad \text{(IV. 19)}$$

where N_w is the number of standing waves for a *sweep wave length* λ_w that would exist between radar and target if a radio signal of frequency W were transmitted. It should be noted that ΔN depends neither upon radio frequency F_o nor upon modulation frequency f_m but is completely determined, for a given propagation velocity c, by sweep width W and range R. Change $\Delta\psi$ in phase of received echo relative to direct mixing signal, caused by modulation, is simply $2\pi\Delta N$ radians.

The output of the radar receiver will vary, during each single sweep of transmitter frequency, in accordance with the envelope or profile of a portion of the standing-wave pattern extending over ΔN cycles, since that much of the pattern moves past the receiving antenna. This is true even though the output may be generated artificially in the receiver, by mixing with the received signal another signal taken directly from the transmitter as described

in section 5 of Chapter III. A minor difference between the receiver-output variation and the resultant-signal amplitude variation at the receiving antenna is that the latter is determined by the range from transmitting antenna directly to receiving antenna and the range from transmitting to receiving antenna by way of reflection at the target, while the receiver output depends rather upon the fixed range equivalent of the antenna lines and mixing-signal path and upon the target range. The output signal may also be regarded as caused by variation, during modulation, of the phase of echo-signal voltage with respect to mixing-signal voltage in the receiver.

To study the way in which errors in indicated range can occur, it is convenient to examine in detail an example of the way in which radar output and counter output vary with range for a special case. The case chosen here is that of a total modulation of 5 per cent of the radio carrier frequency, and of reflection from a plane, perfectly conducting, perfectly stationary target. For the case chosen, F_o is $20W$ and λ_w is $20\lambda_o$. Because it simplifies the situation to do so but does not alter the results in any significant way, variation of signal strength with range will be neglected.

Fig. IV.-11 shows, at (a), the profile or amplitude envelope of the electric-field standing-wave pattern under the special conditions chosen, for ranges up to 15 r-f wave lengths (30 standing waves), or ¾ sweep wave length, from the target. Short vertical strokes mark a set of selected (and numbered) fixed ranges, and about each of these a heavy section indicates the region of the pattern swept or scanned in modulation. At (b) are shown, for the selected ranges, vector diagrams of local mixing (reference) signal e_1, target-echo (received) signal e_2, and their resultant e_3 whose magnitude controls low-frequency radar-output voltage. Section (b) also shows the limits of relative-phase change $\Delta\psi$ resulting from modulation, which correspond to the scanned regions of the standing wave. Section (c) of the figure shows the selected scanned standing-wave segments enlarged, representing the beat-note output wave form of the final detector of the radar either with or without a flat amplifier. Section (d) shows the corresponding output

of the limiter, which changes only during those portions of its input signal (c) lying between the dashed lines.

The type of beat-note amplifier used in altimeters, having a gain which increases with frequency at 6 decibels per octave (that is, per 2:1 increase in frequency), produces as output the derivative with respect to time of the input signal. Differentiated-signal wave forms as produced by such an amplifier are shown in section (e), and limiter output with differentiated input in section (f) of the figure. Each small arrow in section (e) or (f) indicates

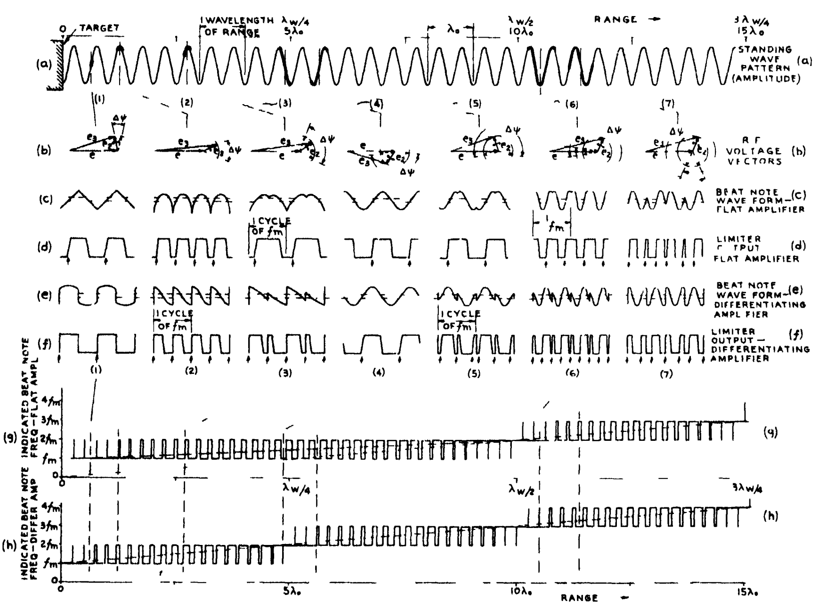


Fig. IV.-11. Occurrence of fixed error, for the case $W = F_o/20$.

occurrence of one count. Sections (g) and (h) show averaged counter output, measuring beat-signal frequency, for signals from a flat amplifier [section (c)] and a high-boosting amplifier [section (e)] respectively.

For extremely short ranges, only a very small fraction of a standing wave is swept over the radar during modulation. The output wave for almost any very short range therefore has the triangular form shown in section (c) for radar position (1), to which corresponds the square wave

from the differentiating amplifier shown in section (e). Sections (d) and (f) show that the corresponding limiter output has the original modulation frequency f_m for either type of amplifier. If, however, the radar is located just at the peak of a standing wave as in position (2), its direct output will have the sharply cusped form shown in section (c) for that position, while the differentiated output has the sawtooth form of the slope of the cusped curve, as shown in section (e). Sections (d) and (f) show that a count is registered twice per modulation cycle for either type of amplifier, so that the counter output for position (2) indicates twice modulation frequency. In sections (g) and (h) is plotted the result that for extreme short ranges the counter indicates frequency f_m independently of range, except that for ranges of just an integral number of quarter r-f wave lengths twice modulation frequency is indicated. Because of the finite time required to reverse the motion of the modulator diaphragm, actual output wave form does not have the perfectly sharp cusps of the figure.

As range increases, the regions of double frequency increase and those of single frequency decrease in width. At a range of 2½ r-f wave lengths (5 complete standing waves) or ⅛ sweep wave length, the pattern motion covers just one quarter of a standing wave ($\frac{1}{8}\lambda_0$). Points at or near integral quarter-wave-length ranges (echo-signal and mixing-signal vectors in line) still correspond to twice-modulation-frequency output and points near intermediate eighth-wave-length ranges (vectors at right angles) correspond to modulation-frequency output, with either amplifier. At such other intermediate points as (3), signal wave forms are as illustrated for that point: the flat-amplifier signal produces an indication of modulation frequency f_m and the differentiated signal an indication of $2f_m$. Investigation of some slightly displaced points indicates that in this region near 2½ wave lengths the flat amplifier gives a double-frequency output for just one fourth of the possible range values, while the differentiating amplifier gives double frequency for just half the possible range values, as shown in sections (g) and (h).

By this time it should be apparent that whenever the portion of the standing-wave pattern moving past the radar as a result of frequency modulation includes one maximum or minimum (echo-signal vector moves once through line of mixing-signal vector), an output signal having a wave form somewhat resembling that at position (3) will be produced. When no extremum is passed (vectors do not align at all during modulation cycle) the output signal will somewhat resemble that for position (1). Cusped signals like those for positions (2) and (3) give rise with a flat amplifier to double-frequency limiter output if both cusps cross as at (2) the region between the dashed lines, which represents the active portion of the limiter-input characteristic (this is usually adjusted to lie at the average value of the signal voltage), or to single-frequency output if only one cusp crosses this region as at (3). On the other hand, appearance of the second cusp at all in the original beat signal requires a reversal of signal slope during each half cycle of modulation and so gives rise to double-frequency output from a differentiating amplifier.

At a range just under 5 r-f wave lengths or ¼ sweep wave length, the pattern motion is just under ¼ r-f wave length or just under ½ standing wave (relative vector-phase rotation just under 180 degrees). Double-cusped signals will result for any range in this region except that of position (4), which is analogous to position (1), so double-frequency differentiated-signal counter output will occur as plotted in section (h) for all such ranges except (4). Single-frequency limiter output will occur for just half the range values in this region with signal from a flat amplifier, as plotted in section (g). At ranges just over 5 r-f wave lengths or 10 standing waves, the range of pattern motion will be just over ½ of one standing wave. No more signal of the type found at positions (1) or (4) will be observed at these or greater ranges (echo-signal vector will then always rotate through line of mixing-signal vector), so no further indication at frequency f_m will be given by a counter driven from a differentiating amplifier. For most range values just above 5 wave lengths, signal of the position-(3) type will occur and the differentiating amplifier will produce a double-frequency count, while the flat amplifier will still produce

double-frequency counting for one half of the possible range values and single-frequency counting for the other half.

For the particular range of position (5) a new type of beat-note wave will now appear, having again two cusps per modulation cycle but showing two reversals of slope instead of one per half cycle (echo-signal vector crosses line of mixing-signal vector twice per modulation sweep). This will produce single-frequency counting with a flat amplifier, but as shown by the wave forms of sections (e) and (f) for position (5) will result in a count of $3f_m$ if the amplifier gain increases 6 decibels per octave with increasing frequency. Thus the counter-output variation with range for this region still proceeds as in sections (g) and (h) of the figure, with section (h) entering a region of new behavior at a range of 5 r-f wave lengths or ¼ sweep wave length.

From 5 to 10 r-f wave lengths, wave forms of the same general type as those at positions (2), (3) and (5) occur repeatedly, with the position (5) type becoming more and more prominent until just below 10 wave lengths the position-(2) or -(3) form, with blunted cusps, appears only at integral-quarter-wave ranges. For all other ranges near 10 wave lengths, the flat amplifier gives rise to counting of $2f_m$ and the differentiating amplifier to $3f_m$.

Just beyond 10 r-f wave lengths or ½ sweep wave length, with pattern motion just over ½ wave length or one whole standing wave, a position such as (6) produces another new wave form, with three reversals of slope per half modulation cycle, while position (7) produces a wave of the sort found at (5) but with both cusps extending across the active region of the limiter. Position (6) gives double-frequency counting with the flat amplifier but quadruple frequency ($4f_m$) with differentiating amplifier, while position (7) gives triple frequency with either amplifier. For the flat amplifier, triple-frequency counting occurs just above 10 wave lengths only at ranges of an odd integral number of eighth wavelengths, with double frequency for all other ranges in this region. For the high-boosting amplifier, quadruple frequency is found in this region at integral-quarter-wave-length ranges and

triple frequency at all other ranges. Above 10 wave lengths, no further single-frequency counting occurs for the flat amplifier nor any double-frequency counting for the differentiating amplifier. Similar considerations for ranges between 10 and 15 r-f wave lengths lead to the remaining values of counter output plotted in sections (g) and (h) of the figure.

Examination of the vector diagrams of section (b) of the figure will show that they follow two simple rules at all ranges. The variation $\Delta\psi$ of relative phase during modulation depends only on the value of range in sweep wave lengths, and is just 4π radians or 720 degrees per sweep wave length. The average phase lag of the echo signal over the modulation cycle depends only on phase change at reflection and on range in average radio-frequency wave lengths, increasing by 4π radians for each r-f wave length increase in range. In the case shown, the echo-signal vector oscillates during modulation about an in-line position for even-eighth-wave-length (r-f) ranges, and about a quadrature position with respect to the mixing signal for odd-eighth-wave-length ranges.

The results of this lengthy consideration of special cases can be stated more concisely, with reference to the counter outputs plotted in sections (g) and (h) of Fig. IV.-11 for various accurately constant ranges. In the example chosen, with sweep width W set at 1/20 of average radio frequency F_0, the fundamental range region found from the special cases extended over 5 wave lengths at F_0. But in general as well as in this example, the fundamental interval is one quarter of the sweep wave length c/W or λ_W, corresponding to the frequency width W of the band swept in modulating.

The graphs indicate that a counter fed directly with beat-note signal from a flat amplifier must indicate either modulation frequency or twice modulation frequency and nothing else for any range from zero to one-half sweep wave length, indicating either twice or three times modulation frequency for ranges from one-half to three-quarters sweep wave length, three or four times modulation frequency for ranges from three-quarters to one sweep wave length and so forth. Similarly, a counter fed from a

differentiating amplifier must indicate either modulation frequency or twice modulation frequency for ranges from zero to one-quarter sweep wave length, either twice or three times modulation frequency for ranges from one-quarter to one-half sweep wave length, and so on. The exact range at which each change from one to the other of the two possible frequencies occurs depends on average r-f wave length λ_o, sweep wave length λ_w, and phase shift on reflection. Fig. IV.-11 applies in detail only to the special modulation ratio and perfectly conducting reflector chosen as an example.

The frequency indicated is, in the case shown the correct one for the actual constant target range only at ranges of exactly one-half, one, 1½, and so on, sweep wave lengths. This happens to be true for both types of amplifier in the example chosen. The frequency counted is never the true range frequency except at ranges that are exact multiples of one-quarter sweep wave length, and is not even always correct for all of those ranges. The change from steady indication of one to steady indication of the other of the two frequencies possible in any range region takes place completely for an extremely slight change in range — perhaps 1/32 of a radio-frequency wave length or less in the example used.

By equation (II.22), a change of f_m in range-beat frequency corresponds to a change of $\frac{1}{4}c/W$ in range. That is, the jump in indicated range between the two counter-output levels possible near any given actual range is just one-quarter sweep wave length. This is the maximum error in indicated range that the effect under discussion can produce [except, as in (g), with a flat amplifier and at ranges less than $\frac{1}{4}\lambda_w$]. This limiting absolute value of systematic error does not vary with actual range, and has often been called the *fixed error* of the radar-indicator system.

Fixed-error phenomena are easily demonstrated and very striking characteristics of frequency-modulated radar under laboratory conditions, where test ranges can be held extremely constant. In the actual use of f-m radar, however, sufficiently constant range to make fixed error evident is highly improbable. Actual ranges will vary either in

coherent fashion because of definite relative motion of radar and target or in random fashion because of surface roughness or other variability of the target. The counter will indicate the average of the various signal frequencies produced by such range variations.

Referring again to Fig. IV.-11, section (g), it may be seen that for ranges near 5 wave lengths of F_0 (¼ sweep wave length) and a flat amplifier, modulation frequency is present for one half of the possible range values and double modulation frequency for the other half. The probability of occurrence of either frequency for a varying range in this region is therefore one half. The average frequency indicated by the counter is correspondingly $(3/2)f_m$.

Consideration of slightly variable ranges of other magnitudes indicates that the smoothed counter output with a flat amplifier, for other than perfectly fixed ranges, varies with range according to the broken line of Fig. IV.-11 (g), and that a similar result holds for section (h). Examination of the dashed line of section (g) and its dotted extension shows that, with a flat amplifier, averaged counter output is directly proportional to range beyond one-half sweep wave length, but that this simple relationship is replaced by a different one for still shorter ranges. From section (h), averaged counter output when fed from a differentiating amplifier is directly proportional to range plus one-quarter sweep wave length for all ranges.

From what has been said, it should be evident that the condition for occurrence of fixed error with a flat amplifier is either that the cusp of the output wave at modulation turn around should cross the active region of the limiter characteristic for a large number of successive modulation cycles, or else that it should fail to cross this region for a large number of successive modulation cycles. Either condition may be met with a highly stable radar working against a target at accurately fixed range. Wobbling transmitter-output phase or modulation sweep, or a moving target, will usually prevent fixed error. If, however, radar and target are steadily approaching head-on at such a rate that range decreases by just one-quarter sweep wave

length per modulation cycle, a fixed-error condition may again exist. This state of affairs requires not only high stability of radar and target but also of the line-of-sight component of relative velocity, and its occurrence for any significant length of time in normal operation is extremely improbable.

The cause of fixed error is the periodic breaking up of the beat-note signal by a phase jump at modulation turn around, coupled with the inability of the combination of limiter and averaging cycle-rate counter to measure fractions of cycles. Fixed error is therefore a characteristic of the use of f-m radar under conditions for which range-beat frequency exceeds speed-beat frequency. For cases in which speed frequency exceeds range frequency (which may even be zero) and a flat amplifier is used, fixed error does not exist. In the former case, the continuously acting averaging pulse counter indicates range; in the latter case, it indicates speed.

The space here given to the discussion of fixed error is entirely out of proportion to the practical importance of the phenomenon, but not at all out of proportion either to its nuisance value or to its fundamental importance for the understanding of f-m radar. This discussion is felt to be justified because any worker in the f-m radar field will almost inevitably encounter fixed error in some form, and because it indicates the necessity of using a differentiating amplifier if linear range measurement is to be extended down to extremely short range.

3. SWITCHED COUNTERS

a. *Speed Counter.* As described in section 2c above, a counter can be made to produce either a positive or a negative average-current output, directly proportional to the frequency of the counter-input signal. Counting can be stopped at will by opening either the charging or the discharging circuit of the counter capacitor. If the counter diodes are replaced by triodes, either circuit may be opened by biasing the proper triode beyond plate-current cut off.

Let a positive-output counter and a negative-output counter both be energized by a common limiter, both feed a common load, and both be linearized by a common cathode

follower. Further, let each counter include one diode in its capacitor-discharging circuit and one triode in its charging circuit, with the triodes alternately cut off for equal time intervals by a symmetrical square wave of voltage applied in push-pull to their grids, as in Fig. IV.-12. If now a signal of constant frequency is

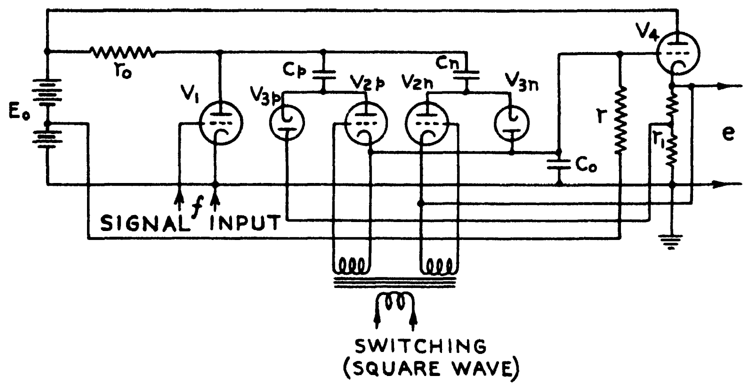


Fig. IV.-12. Switched counter.

applied to the limiter and the two counter capacitors C_p and C_n are equal, then the total positive charge fed to smoothing capacitor C_0 and load resistor r, while the negative counter is inactivated by cut-off bias on triode V_{2n}, will equal the negative charge supplied while the positive counter is inactivated by cutting off V_{2p}. The net charge delivered by the counters over a complete switching cycle will therefore be zero and the average voltage developed across r will likewise be zero.

The instantaneous voltage across C_0 and r will not be zero, however. It will increase in steps as successive increments of positive charge, one for each cycle of the signal being measured, are supplied during the intervals of activity of the positive counter. During the intervening intervals of activity of the negative counter, the voltage across C_0 will decrease in steps as negative charge increments are supplied. Between steps there will be steady voltage changes because of slight leakage of charge to or from C_0 through r.

The cathode of follower tube V_4, to which the charging tube of the negative-output counter is returned directly,

operates at a positive voltage with respect to the follower-tube grid, to which the discharging tube of the negative-output counter and the charging tube of the positive-output counter are both connected, while the cathode-resistor tap, to which the discharging tube of the positive-output counter is returned, operates at a voltage negative of the follower grid. In order that the average counter output over the switching cycle may balance to zero independently of the limiter-output voltage swing, the cathode-resistor tap must be negative of the follower grid by as much as the follower cathode is positive of its grid. With equal internal voltages in the counter tubes, this ensures equal effective limiter swings E_o' [see equation (IV.6)] for both counters.

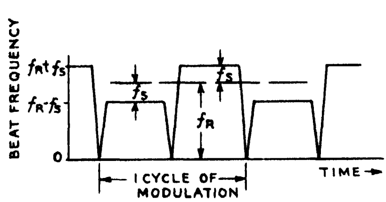


Fig. IV.-13. Frequency variation of radar output.

Suppose now that this switched-counter system is fed with the beat-note output of an f-m radar, which varies in frequency as shown in Fig. IV.-13 (see Chapter II., section 4f); it is switched in synchronism with the radar modulation, so that the positive counter is active during the upward modulation sweep of the transmitted radio frequency, when beat-note frequency is $f_R - f_s$. The negative counter then operates during modulation downsweep, with beat-note frequency $f_R + f_s$. The charge contribution due to range frequency, which is the same throughout the switching cycle, will average out and leave the net counter output independent of range. During downsweep the presence of a speed frequency will increase the negative charge delivered to C_o, while during upsweep the positive charge delivered will be decreased by an amount proportional to the speed frequency. A net negative charge proportional to speed will therefore accumulate over the switching cycle. That is, this speed counter is a negative-output device. In the steady state, the average voltage e across r will reach such a value that the resulting current through r will just

equal the average rate of delivery of charge by the counter and so will just carry off this charge.

During upsweep, the positive counter delivers average current

$$i_u = (f_R - f_s) C_s E'_o \tag{IV.20}$$

and during downsweep the negative counter delivers average current

$$i_d = -(f_R + f_s) C_s E'_o, \tag{IV.21}$$

calling the equal counter capacitors C_p and C_n both C_s. Each counter operates half the time, so that the net current is the average of i_u and i_d, while the average output voltage e_s over the modulation (and counter-switching) cycle is

$$e_s = - f_s r C_s E'_o . \tag{IV.22}$$

Since this output is negative and proportional to radar speed frequency, such a switched counter may be called a negative-output *speed counter* and characterized by a *speed-counter sensitivity* h_s, in volts per cycle per second, where

$$h_s = r C_s E'_o . \tag{IV.23}$$

The instantaneous voltage across r and C_o varies with time as shown in Fig. IV.-14, becoming one step more positive with the charge increment delivered on each beat-note

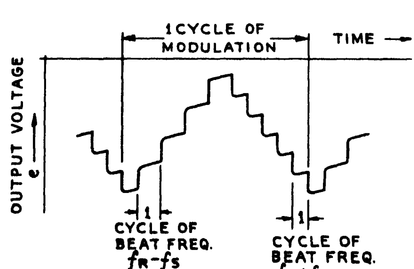


Fig. IV.-14. Wave form of output from speed counter.

cycle during upsweep and one step more negative for each cycle during downsweep. Between steps there is a relatively slow change to more positive voltage due to leakage through r, which must in the steady state be just sufficient to dispose of the excess negative charge delivered by the counter

during the complete modulation cycle. The linearizing cathode follower shown in the circuit of Fig. IV.-12 is necessary to insure that the size of each charge increment will be independent of frequency. Were the effective limiter swing changed by building up of output voltage during the upsweep counting, the excess negative charge delivered to C_o during the complete modulation cycle would depend upon both range and speed. The speed counter would then fail of its purpose.

The above discussion applies to the case of radar range frequency in excess of speed frequency. For the case of speed frequency exceeding range frequency, conditions are reversed and the symmetrical switched counter becomes a range counter with output independent of speed. In either type of operation, the frequency to be cancelled over the modulation cycle exceeds, perhaps by many times, the frequency to be indicated. The balance of the two counters required for proper operation may therefore be rather critical and small fortuitous variations in such parameters as switching phase may produce significant fluctuations in counter output. Because of the balance requirement, capacitance change is not a convenient means of controlling speed-counter sensitivity h_s, but either load resistance r or effective limiter-voltage swing E_o' may be varied to provide such control.

b. *Combined Range and Speed Counter.* Another useful type of switched counter is obtained by using unequal values of capacitance for the positive and negative counters. Using the positive counter during modulation upsweep and the negative counter during downsweep, the upsweep current for radar operation with range frequency exceeding speed frequency is

$$i_u = (f_R - f_s) C_p E_o' \tag{IV.24}$$

and the downsweep current

$$i_d = -(f_R + f_s) C_n E_o'. \tag{IV.25}$$

The average of the counter-output voltage over the modulation cycle is therefore, since each of the above currents flows for half the time,

$$e = \tfrac{1}{2} f_R r (C_p - C_n) E_o' - \tfrac{1}{2} f_s r (C_p + C_n) E_o'. \tag{IV.26}$$

This counter is thus sensitive to both range and speed, with range sensitivity

$$h_R = \tfrac{1}{2} r(C_p - C_n) E'_o \tag{IV.27}$$

volts per cycle per second and speed sensitivity

$$h_s = \tfrac{1}{2} r(C_p + C_n) E'_o \tag{IV.28}$$

volts per cycle per second.

Range sensitivity is positive if C_p exceeds C_n and negative if the reverse is true. Negative-output speed counting occurs for operation as described, but speed output would be positive if the positive counter were used on the downsweep and the negative counter on the upsweep. Subject to the condition that the absolute value of speed sensitivity, (capacitance sum) must obviously exceed the absolute value of range sensitivity (capacitance difference), any desired combination of positive or negative sensitivities may be obtained by proper choice of operating sequence and capacitance values. Independent adjustment of range and speed sensitivities of an unsymmetrical switched counter is inconvenient once they are chosen, however, because interlocking adjustments of both capacitors are necessary.

For radar operation with speed frequency exceeding range frequency, the right-hand sides of equations (IV.27) and (IV.28) are simply interchanged to give range sensitivity proportional to capacitor sum and speed sensitivity to capacitor difference. Again, almost completely free choice of relative sensitivity values is obtainable.

For a counter-load resistor returned to supply or bias voltage e_o rather than to ground, synchronously switched linear counters are in principle capable of producing an overall time-averaged voltage output

$$e = h_R f_R - h_s f_s + e_o \tag{IV.29}$$

with sensitivities h_R and h_s independently open to choice over a wide range of values. A linear counting system is necessary to ensure this simple form of output variation. Because diode biases for positive and negative counters obtained from the cathode circuit of the linearizing cathode follower vary differently with e, however, critical

accuracy is possible only over a narrow range of output voltage.

c. *Extra Counts.* Certain spurious signals are produced by counter switching. In the analysis of this effect to follow, it should be remembered that, conduction in the switching tubes permitting, both counter capacitors are charged on each upswing and discharged on each downswing of the limiter plate voltage. Charging of the positive-counter capacitor or discharging of the negative-counter capacitor registers a count of corresponding polarity. Neither discharging of the positive-counter capacitor nor charging of the negative-counter capacitor produces any count; these are merely resetting operations.

Suppose the limiter to be conducting, and input voltage to the counter capacitors therefore to be at its minimum value, at the instant of switching off the negative counter and switching on the positive counter. The negative-counter capacitor C_n will already be fully discharged, having just completed a count, and the opening of its charging circuit by the switching will have no effect other than to prevent further counting. The positive-counter capacitor C_p will already be discharged also, and the closing of its charging circuit will have no effect except to prepare for a normal count on the next positive limiter swing.

Similar conditions hold when switching off the positive counter and switching on the negative counter with limiter plate at minimum voltage. C_p will be fully discharged following its last previous count and will merely be prevented from registering the next count by the switching process which opens its charging circuit. C_n will be fully discharged and closing its charging circuit will merely prepare it to be charged on the next positive limiter swing, in order that it may produce a negative count on the following negative limiter swing.

Switching from positive to negative counter with limiter tube cut off and supplying maximum voltage to the counter again yields nothing remarkable. C_p will be fully charged following a count and nothing will happen as a result of opening its charging circuit. C_n will be discharged and will charge immediately when its charging circuit is closed

by switching, in preparation to count on the next downward swing of the limiter plate.

Upon switching off the negative counter and switching on the positive counter while the limiter is cut off and its plate most positive, the action is rather different. The negative-counter capacitor is fully charged and no immediate action results from opening of its charging circuit by the switching, but a negative count will occur on the next limiter-plate downswing even though the negative counter is then turned off and can thereafter produce no further counts. The positive-counter capacitor is discharged and will charge immediately, registering a count, when the switching action closes its charging circuit. Thus, one extra positive count and one extra negative count result, in this limiter-phase condition only, from the occurrence of switching.

Switching phase and signal phase are unrelated, so switching from negative to positive counter may be expected to occur half the time with limiter conducting and half the time during limiter cut off. There will therefore be on the average one-half extra positive count and one-half extra negative count per complete switching cycle. The symmetrical switched counter or speed counter will show no net error due to switching, since its positive and negative counts are of equal magnitude. The unsymmetrical counter will show an average error per switching cycle of one-half the difference in magnitude of one positive and one negative count.

Extra counts due to fixed error will of course be present in addition to those caused by switching. In a continuously operating counter driven by a differentiating amplifier, an average of one extra count per complete modulation cycle was found in section 2g above to occur (counting at modulation frequency near zero range, for example). This may be regarded as an average of one-half count per unidirectional sweep of radio frequency. With a synchronously switched counter driven by a differentiated f-m radar signal, then, an average extra count of one-half the magnitude difference between one positive and one negative count per complete modulation cycle may be expected as a manifestation of fixed error. The total spurious count due both to switching and to fixed error is thus just the difference between one

positive and one negative count per full cycle of the common modulating and switching frequency.

d. *Phase Lag*. With counter switching accomplished by the same square-wave signal which provides the pulses to reverse the motion of a vibrating modulator diaphragm, there is a phase lag between switching and frequency modulation. This occurs because reversal of diaphragm motion is somewhat gradual and requires the full duration of the driving pulse to be completed, while counter switching takes place immediately upon the rise or fall of the controlling square wave. The result of such modulation-phase lag is that the upsweep counter will be active during a portion of the modulation downsweep and vice versa.

For simplicity, suppose that diaphragm reversal takes place suddenly but is delayed by a fraction δ of the full modulation cycle, and neglect as usual the fraction of a modulation cycle taken by the radio signal to reach the target and return. The total charge delivered by the positive or upsweep counter will then be

$$q_p = [(f_R - f_S) C_p E'_o (\tfrac{1}{2} - \delta) + (f_R + f_S) C_p E'_o \delta]/f_m \tag{IV.30}$$

and that delivered by the negative or downsweep counter will be

$$q_n = [-(f_R + f_S) C_n E'_o (\tfrac{1}{2} - \delta) - (f_R - f_S) C_n E'_o \delta]/f_m \,. \tag{IV.31}$$

The average output e due to leakage through load resistor r of the net charge supplied during the total modulation period $1/f_m$ will be

$$e = (q_p + q_n) f_m r \tag{IV.32}$$

or

$$e = \tfrac{1}{2} f_R r (C_p - C_n) E'_o - \tfrac{1}{2} f_S r (C_p + C_n)(1 - 4\delta) E'_o \,. \tag{IV.32a}$$

Comparison of (IV.32a) and (IV.26) shows that the lag in modulation phase has not altered the range sensitivity but has reduced the negative speed sensitivity of the counter by a factor $(1 - 4\delta)$, to

$$h'_S = \tfrac{1}{2}(1 - 4\delta) r (C_p + C_n) E'_o. \tag{IV.33}$$

Practically, the turn-around lag may be for example 5 per cent of the modulation cycle, resulting in a speed sensi-

tivity only 80 per cent as great as that of an ideal system.

e. *Blanking.* Beside the modulation-phase lag at turn around, strong transient signals may be produced at the same time in special f-m radar systems that use antenna-lobe switching to obtain target-azimuth data. Disturbance of switched-counter output either by phase lag or by synchronous transients may be prevented by interrupting or "blanking" counter operation entirely for a brief interval at each modulation turn around.

Blanking may be accomplished by providing a separate square-wave switching voltage to each counter triode, with the positive or "on" portion of each cycle of each switching voltage arranged to have shorter duration than its negative or "off" portion. The two voltages should of course be so phased that the positive counter acts only during the undisturbed and linear portion of the frequency-modulation upsweep and the negative counter during the similar portion of the downsweep. Disturbance of counter output by non-linearity of frequency sweep at turn around and by the time lag of received-signal turn around due to radio-transmission delay are also eliminated by blanking.

If each counter is turned on only for a fraction $\frac{1}{2}-\gamma$ of the full modulation cycle, both range sensitivity and speed sensitivity will be reduced to a fraction $1-2\gamma$ of the values (IV.27) and (IV.28) found for a switched counter operating under ideal conditions. Stability of the blanking fraction γ is therefore essential to accurate and reliable counter operation.

The effect of possible dissymmetry of switching must also be considered. Suppose the positive counter to be active for a fraction $\frac{1}{2}-\gamma+\epsilon$ and the negative counter for a fraction $\frac{1}{2}-\gamma-\epsilon$ of each modulation cycle, the positive counter operating during radio-frequency upsweep and the negative counter during downsweep. Then, since the net charge delivered by the counter during the modulation cycle must leak off through load r, the average output voltage e may be determined and the modified range and negative-output speed sensitivities are found to be

$$h_R'' = \tfrac{1}{2}(1-2\gamma)\, r\,(C_p - C_n)\, E_o' + \epsilon r (C_p + C_n)\, E_o' \qquad \text{(IV.34)}$$

and

$$h_s'' = \tfrac{1}{2}(1-2\gamma)\, r(C_p + C_n)E_o' + \epsilon r(C_p - C_n)E_o', \qquad \text{(IV.35)}$$

respectively.

Since $C_p + C_n$ is always greater than $C_p - C_n$, and may be many times greater, the effect of switching dissymmetry on speed sensitivity is not extremely great and may (as in the balanced speed counter with C_p and C_n equal) be negligible. Switching dissymmetry ϵ clearly has an exaggerated effect on the range sensitivity of the switched counter, however, and if C_p and C_n are nearly equal may cause serious trouble. This is also true in the case of continuous switched-counter operation without blanking, for which (IV.34) and (IV.35) still hold but with $\gamma = 0$. Extreme stability of switching symmetry is therefore essential to reliable operation. In the case of radar operation with speed-beat frequency in excess of range-beat frequency, corresponding conclusions may be drawn but the roles of range and speed are then interchanged.

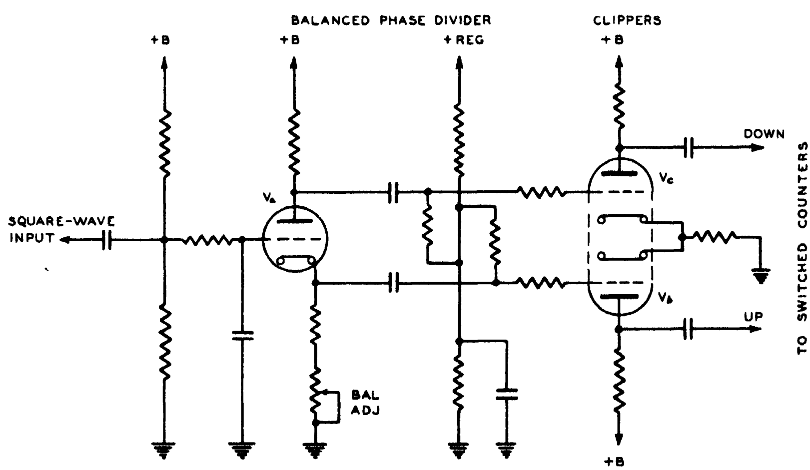


Fig. IV.-15. Counter blanking circuit.

A typical circuit for blanked switching of counters is shown in Fig. IV.-15; wave forms appearing at various points of the switching circuit are shown in Fig. IV.-16. The original constant-amplitude square-wave switching signal with the wave form of Fig. IV.-16(b) is partially integrated by the resistance-capacitance switching-input

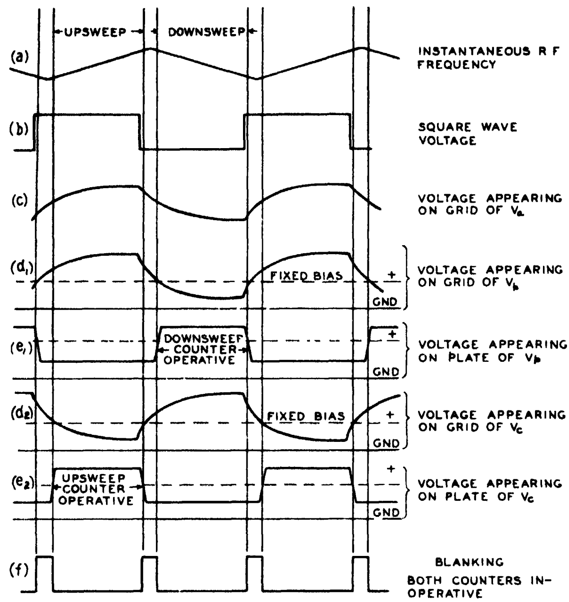


Fig. IV.-16. Development of switching voltages for blanked counter.

circuit, and the resulting modified signal, shown as (c) of the figure, is applied to the symmetrical-output amplifier V_a. Wave forms of the two outputs of V_a are shown at (d_1) and (d_2) of Fig. IV.-16, and each is coupled to the grid of one of the limiter tubes V_b and V_c.

In the absence of switching signal, the grids of both V_b and V_c are maintained at the incipient grid-current point by positive bias applied to high-resistance grid leaks, holding both limiter plates at minimum voltage. Negative swings of the switching signals alternately carry each limiter grid to and beyond cut off, permitting the corresponding plate to rise to supply voltage. The grid-voltage region of limiter activity is indicated by the dashed lines across graphs (d_1) and (d_2) of Fig. IV.-16, while graphs (e_1) and (e_2) show the corresponding limiter-plate voltage wave forms. During the time intervals marked off on all the graphs by the pairs of thin vertical lines, both switching limiters remain conducting and their plates are both at minimum voltage.

Since the output of one switching limiter drives the

grid of each counter-capacitor charging triode, both counters will be rendered inoperative during the intervals indicated in graph (*f*), in which both limiters are conducting. The dashed lines across graphs (e_1) and (e_2) indicate the threshold of counter operation: for limiter plate voltages above the dashed line, the corresponding counter is operative, while for voltages below the line it is inoperative. Graph (a), to the same time scale as the others, shows the variation of transmitted radio frequency.

Symmetry of switching depends upon output balance of amplifier V_a, which may be controlled by adjustment of its plate or cathode load resistor. Blanking fraction γ depends upon switching-limiter bias and original square-wave amplitude, which must therefore both be stable. Stability of switching is essential to accurate switched-counter operation, and the additional possibilities for instability introduced by the inclusion of blanking circuits must not be overlooked. The circuit shown has appeared in practice, however, to be sufficiently stable not to decrease the counter accuracy too seriously.

By separating the switching off of one counter from the switching on of the other, blanking introduces further possibilities for phase of limiter-output signal at instants of switching, in addition to those enumerated in section 3c above. Examination of all such possibilities and their probabilities of occurrence, however, again leads to the conclusion that on the average an extra count of one-half the difference in magnitude between one positive and one negative count per modulation cycle results from counter switching. Fixed error can be more pronounced in a switched counter than in a continuously acting one, by upsetting the partial cancellation of the two counter outputs, and the use of blanking does not seem in practice to alter significantly the fixed-error effects.

f. *Effect of Signal Level.* Unless the amplitude of the signal to be measured is sufficient to swing the limiter plate current from cut off to saturation, no counter of the type described can give accurate results. So long as the signal amplitude is adequate to ensure that the limiter plate current remains cut off during each signal cycle for a sufficient period to permit the counter capacitor to become

fully charged, even at the highest frequency to be measured, and so long as the plate current remains saturated on each cycle long enough to permit complete capacitor discharge, full counter accuracy may be expected for any interference-free signal.

The presence of noise or other interference complicates the situation. Fig. IV.-17 shows how noise superposed on a signal may give rise to extra counts, even though the signal may considerably exceed the noise in amplitude. Noise cannot reduce the count so its effect does not average out. The magnitude of the error produced in this way and its dependence on signal/noise ratio have not yet been

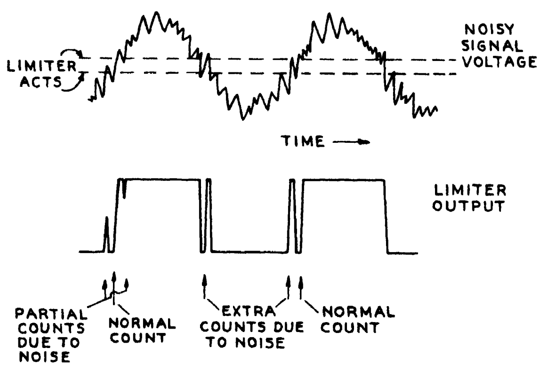


Fig. IV.-17. Effect of noise on counter action.

accorded the full investigation that the subject deserves. Rules of thumb have been worked out, however, for practical uses in which errors up to a few per cent are tolerable. For these cases, sufficient amplification should be provided so that the noise alone is approximately sufficient to swing the limiter between cut-off and saturation. Then signals with amplitude greater than 10 times the r-m-s noise amplitude are adequate for satisfactory operation, if a differentiating amplifier is used. With a flat amplifier, the noise would be of lower effective frequency and a smaller signal/noise ratio might be permissible.

The great increase in minimum acceptable signal/noise ratio required by this automatic measuring equipment over that necessary for a skilled operator just to perceive an oscilloscope signal by eye, or a communication signal by

ear, is to be noted. It is the price paid for elimination of human judgment and for quantitative accuracy.

Square-wave input signal to the counter must be at such a high level that any change in effective input swing E_o' due to variation of internal potentials (electron-emission velocities and contact potentials) of the diodes can not introduce an excessive percentage error in counter sensitivity. Internal-potential changes in the several diodes of a switched counter may add up to produce a total change in effective counter-input amplitude of as much as one volt. Counter-output level also must be made sufficiently high to ensure accurate operation of subsequent circuits, a requirement which determines the maximum frequency that can be counted while using a limiter tube of any given current-handling capacity. In practice, it has not been difficult to secure sufficient counter output for satisfactory operation under conditions required by f-m radar applications, with beat-note frequencies in the range from a few hundred cycles to a few tens of kilocycles per second.

4. FREQUENCY-SELECTIVE CIRCUITS

a. *Single Target.* Instead of averaging counters, passive circuits having a response dependent on signal frequency may be used for frequency measurement. Such circuits have not been fully investigated for single-target f-m radar operation, because it was felt that counters might more easily be made stably accurate and the urgency of war-time development required a choice between the two lines of attack.

For radar use, a selective circuit for frequency indication, or "discriminator", must produce an output varying monotonically, substantially, and preferably linearly with signal frequency over a very wide frequency range. Substantial monotonic variation with frequency is produced for example by an integrating circuit fed with square-wave signal of constant amplitude. Such a circuit provides a triangular-wave output with amplitude inversely proportional to frequency. A differentiating circuit fed with constant-amplitude sine-wave input signal produces a sine-wave output with amplitude varying linearly with frequency.

Purely sinusoidal input signal is necessary for use of the latter circuit, but is not readily provided over the required frequency range in the presence of noise.

The integrating circuit can be made to produce an output wave of peak amplitude decreasing linearly with increasing frequency over a 10:1 range of frequencies by bridging its input with a series-resonant circuit tuned to a high frequency and suitably damped. The output voltage and therefore the input frequency may be measured by a peak-reading vacuum-tube voltmeter. Such a system has been used successfully in reception of facsimile signals which employ a frequency-modulated sub-carrier. Switched operation may be obtained by use of two grid-controlled voltmeter rectifiers.

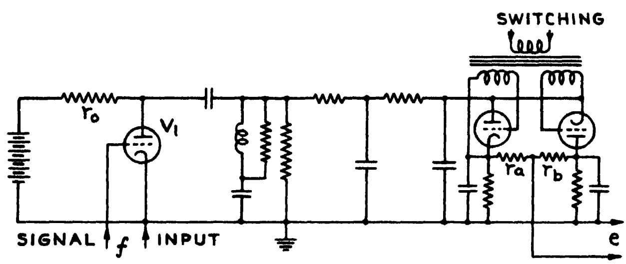


Fig. IV.-18. Discriminator circuit for frequency measurement.

A circuit of this sort, shown in Fig. IV.-18, was given limited trial[2] under f-m radar operating conditions. Relative range and speed sensitivities of this circuit are controlled by choice of resistors r_a and r_b in the output-voltage divider. As a continuously acting device, using a single rectifier and no switching, satisfactory operation was obtained. But when operated as a switched system to measure a combination of range and speed, as shown in the figure, the maximum output obtainable was less than one tenth of that found necessary for satisfactory operation. This limitation could only be overcome by use of excessively high limiter-plate voltages, with a correspondingly highly rated limiter tube.

For cases in which its low output is acceptable, the above frequency-measuring device appears to offer a possibility of eliminating fixed-error effects. Provided several beat-note cycles occur per half sweep of the radar fre-

quency modulation, the reading of each peak voltmeter may be fully determined by the discriminator-output level during the highly linear or steady-state middle portion of a frequency sweep. Then, by blanking the voltmeter rectifiers, it should be possible to prevent the disturbances taking place near modulation turn around from affecting the voltmeter outputs at all. Such an arrangement should be insensitive to exact phase of blanking and should, of course, eliminate effects of modulator phase lag as well as of fixed error.

Because discriminator output is insufficient to make it useful in those applications which were of main interest, the blanked-peak-voltmeter arrangement was not tried. In the case of the simpler switched-voltmeter arrangement of Fig. IV.-18 which was tried, however, fixed-error effects appeared under limited investigation to be of the same order as those found in the use of counters.

The switched discriminator-voltmeter combination requires two less tubes (a twin diode and a cathode follower) than the switched counter, but pays for this simplification by restricted frequency range, much lower output, and dependence of its calibration on the values of a considerably larger number of circuit elements. By sacrificing simplicity, a blanked discriminator-voltmeter system may perhaps be freed of modulation-phase and fixed-error disturbances. Under some conditions, filtering time constants may be lower for the discriminator than for the counter.

Quite a different application of a selective circuit permits determination of both range and speed without synchronous switching. Either a single counter or a single discriminator-rectifier fed with the f-m radar signal from a moving target will produce an output current varying at modulation frequency as shown by Fig. IV.-13. In a resistive load this current will produce an average output proportional to range. In a resonant load circuit tuned to modulation frequency, it will produce an alternating output of amplitude proportional to speed, which may in turn be rectified for speed measurement. As in the switched counter, linear variation of average direct output current with input frequency is necessary in the system driving the resonant circuit, to avoid dependence on range of the speed sensitivity of this speed-measuring device.

Trials of speed measurement by amplitude of frequency modulation of radar beat note (see Fig. IV.-13) were made only with non-linear frequency modulation of the transmitted radar signal. A simple counter was used with a parallel-tuned load circuit in series with the normal load r, C_0 (see Fig. IV.-3). Strong modulation-frequency output was obtained from the tuned circuit but was found not to be dependent only on speed. A major part of this output was found to be due to dissimilarity of radar frequency-modulation wave form in the vicinity of the two modulation turn-around points, which altered the frequency variation of the radar beat note so as to give a component at modulation frequency even with a stationary target. With sufficiently linear modulation this trouble should be absent and the modulation-frequency resonant speed detector should be a useful device. Unfortunately, no tests were made under such conditions.

b. *Multiple Targets.* Multiple-target operation requires signal-utilization apparatus that will indicate for each value of range whether or not a target is present. In f-m radar technique, a beat-frequency resonant circuit performs the same function as does a range gate in pulse radar. Presence or absence of a target at a given range is determined by applying the radar output to a resonant gate tuned to pass signal corresponding to that range only, and indicating whether output from the gate is present.

Various range values may be explored for targets by using

(1) Multiple fixed-tuned gates, each set to test a different range.

(2) A single variably tuned gate which may be made to scan the radar output-frequency spectrum for targets.

(3) A single fixed-tuned gate over which the radar output spectrum may be scanned by varying the rate of change of radio frequency used in frequency modulation of the radar.

Multiple-gate indicators are complex but deliver information rapidly. Scanning single-gate indicators are simpler but deliver information quite slowly, since it is necessary

at each range value to allow time for the response of the resonant gate to build up.

An example of a multiple-gate system which is its own indicator is the resonant-reed assembly of a Frahm frequency meter. This device might be but has not been developed as an f-m radar indicator. It would be quite suitable for indicating target distribution along a single line of search, but less suitable for panoramic display of the varying target distribution seen by a radar scanning in azimuth.

An example of a variable tuned scanning gate is the well known heterodyne wave analyzer, in which an actual fixed-tuned circuit is made by artifice to behave as if variably tuned. For radar indication, the output of such a wave analyzer, scanning periodically by variation of heterodyne-oscillator frequency, might be applied to deflect or brighten the light spot of a cathode-ray oscilloscope. The oscilloscope would be made to scan a range scale in synchronism with the frequency scan of the analyzer.

For a radar with scanning by periodic variation of range sensitivity, the output of a single fixed-tuned gate may be used to brighten or to deflect horizontally the light spot of a cathode-ray oscilloscope, with the spot made to scan vertically in synchronism with the radar-sensitivity scan. Trial of such a system has shown it to operate as expected. It has of course been found that accurately uniform sweeping of transmitter frequency during modulation, hence accurate constancy of beat-note frequency for any fixed range, is essential to avoid degradation of range resolution by apparent spreading out of targets. Linear detection of the radar signals, and indeed linear response throughout the system, is also essential to avoid inter-modulation of signals from various targets and consequent indication of spurious or ghost targets. If fine range differences are to be indicated while searching in azimuth, operation necessarily becomes very slow.

5. UTILIZATION OF COUNTER OUTPUT

a. *Meter Indication.* F-m radar information is finally utilized most simply, at least for the case of a single target, by applying the output from a counter such as that

of Fig. IV.-2 to a current-measuring instrument, which in turn gives an indication of target range. Meter current increases linearly with increasing range. The maximum meter current that may be supplied directly by the counter without loss of accuracy is determined by the properties of the limiter tube.

Referring to Fig. IV.-7, it will be seen that a given type of limiter pentode can supply only a definite maximum plate current if dependence of this current on grid voltage must fall to zero while the grid is still negative, as is necessary for sharp limiting. The minimum permissible plate-feed resistor r_o is determined by this maximum available current I and the limiter plate-voltage swing between plate-current saturation and cut off. Neglecting the small diode potentials and biases, this is the same as the effective charging-voltage swing E_o' applied to the counter, and

$$r_{o_{min}} = E_o'/I \tag{IV.36}$$

Linear operation accurate to one per cent requires that charging of the counter capacitor continue for an interval of at least four times the time constant of the charging circuit, even at the highest frequency to be measured. Since charging takes place during one half of each beat-note cycle only, while limiter plate current is cut off, the charging time constant must never exceed one eighth of the beat-note period. Therefore

$$r_o C \leqslant 1/(8f_R), \tag{IV.37}$$

or

$$C_{max} = I(8f_{R_{max}} E_o'). \tag{IV.38}$$

Using this limiting capacitance in equation (IV.2), the maximum current-output capability of the counter for accurate operation is

$$I_{max} = \tfrac{1}{8} I. \tag{IV.39}$$

For the 6SH7, a useful limiter tube, the maximum negative-grid plate-saturation current is about 5 milliamperes and the maximum linear counter output about 625 microamperes.

Where a more rugged instrument than the counter can actuate directly is required, or where heavy meter shunting is necessary to supply electro-mechanical damping, the meter may be placed in the cathode circuit of a counter-linearizing cathode follower (see Fig. IV.-5). This arrangement, used in altimeters, also permits introduction of controlled non-linearity in the relation between frequency counted and meter current. Of course these meter arrangements using an unswitched counter, described above as range indicators for the case of radar range-beat frequency in excess of speed-beat frequency, will indicate speed instead of range if speed frequency exceeds range frequency.

b. *Limit Relays.* Instead of giving a continuous meter indication of range, speed, or a combination of both, arrangements may be made to actuate a relay when the counter output reaches, or departs to a predetermined extent from, a preset value. For example, it is desired in certain cases to release a missile when a radar-bearing aircraft reaches a range from the target determined by closing speed and other factors. Prior to release, range and counter output exceed a predetermined limiting value. A relay must be actuated when net switched-counter output due to both range (counted positively) and speed (counted negatively) falls to that limiting value.

The desired result may be accomplished by connecting the counter-follower cathode directly to the cathode of a relay-actuating tube and returning the grid of the latter to the preset positive voltage at which release is required, as shown in Fig. IV.-19. For counter-output voltages exceeding the voltage difference between the voltage-divider taps to which counter load and relay-tube grid are respectively returned, relay tube V_5 will be cut off by the positive voltage applied to its cathode. When counter-output and bias-difference voltages become substantially equal, V_5 begins to conduct and the relay is actuated. For still lower counter output, cathode follower V_4 is cut off and the counter is no longer linear; this is unimportant, as the equipment has then served its purpose and need operate no longer. Radar-range and speed-counter sensitivities, and voltage difference between cathode-follower

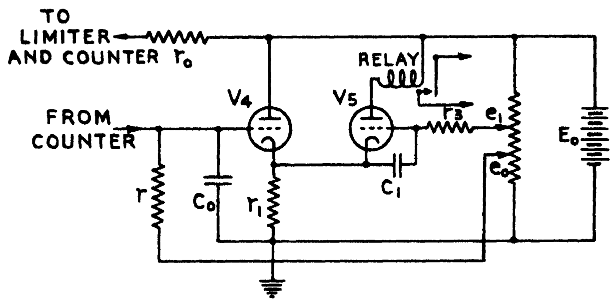


Fig. IV.-19. Circuit for relay operation.

and relay-amplifier biases, are all preset to ensure release under the desired condition.

The relay tube is biased through resistor r_3 and its grid is bypassed to cathode by capacitor C_1 to filter out the modulation and beat-note frequencies of Fig. IV.-14, which are present on the cathode of follower V_4. The grid-cathode circuit of the relay tube is thus controlled by a hum-free signal and smooth operation is assured, at the cost of a slight increase in time lag over that already produced by the counter-load time constant rC_0.

A different requirement exists where it is desired to maintain a constant predetermined range, as in the use of a radar altimeter to maintain constant aircraft clearance over the underlying terrain. This requirement has been met by use of the null-type counter of Fig. IV.-6, with the range to be maintained determined by the setting made on the calibrated voltage divider to which the counter load is returned. Departure of voltage of output point A from zero, corresponding to departure of range from the preset value, must then actuate a polarized indicator. Point A is therefore connected to the grid of a relay-actuating tube which has two relays in series in its plate circuit. One of these relays operates at a low current value and the other at a considerably higher current; the actuating tube has its cathode biased to such a voltage that its plate current with grid at zero lies between the operating values for the two relays.

Normally, range is at the preset value, point A is at zero voltage, one relay is actuated, and the other relay is not. The relay contacts are so connected to signal

lamps as to give a "Range Correct" indication in that condition. For greater range, increased counter output will make point *A* negative, decreasing the relay current so that neither relay is actuated and a "Range Excessive" signal is given. For a smaller range than the value preset for balance, voltage at *A* is positive, both relays are actuated and a "Range Insufficient" signal is given. If the radar operates with speed-beat frequency in excess of range frequency, the same device will serve to indicate departure of target speed from a chosen value.

c. *Servo Systems.* Still another possibility that has found useful application is the utilization of counter output to actuate a servo mechanism, which in turn operates to maintain net output constant. In order to achieve this, the servo mechanism may in principle control any one or more of the following variables:

(1) Range from radar to target.
(2) Speed of radar relative to target.
(3) Radar range sensitivity.
(4) Radar speed sensitivity.
(5) Range-counter sensitivity.
(6) Speed-counter sensitivity.
(7) Bias voltage to counter load.

The further possibility of allowing net counter output to vary, and using a servo mechanism merely to measure this variation and to produce a mechanical motion related thereto, involves little that is peculiar to f-m radar and so will receive little discussion here. Likewise, the general theory of servo-mechanism response and stability is accessible elsewhere[3] and so need not be considered extensively here.

Maintenance at zero of the output of a null-type counter by a servo adjustment of the voltage-divider tap is easily accomplished and has proved a useful means for developing a mechanical motion proportional to radar range. In the two-relay arrangement described above for limit indication, it is only necessary to have actuation of the high-current relay start a reversible motor, geared to the potentiometer r_a in the direction to reduce the voltage at the tap, while

opening of the low-current relay starts this motor running in the direction to increase tap voltage as in Fig. IV.-20.

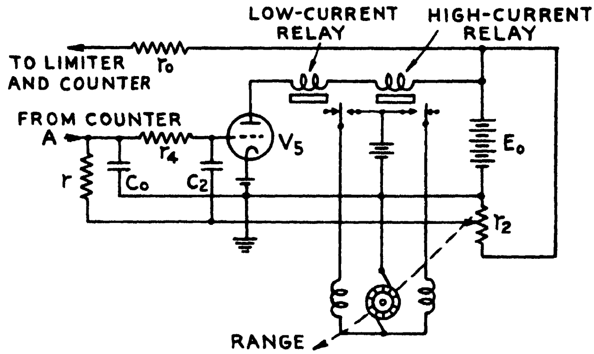


Fig. IV.-20. Simple range servo.

When voltage at counter-output point A is zero, the low-current relay will be actuated and the high-current relay open, and the motor will stop. The relation between servo-produced shaft rotation and radar range will be the same as the relation between shaft rotation and divider ratio for which the potentiometer r_2 has been constructed.

In order for this simple servo system to be stable, it is necessary that there be for any given radar range a finite region of potentiometer settings, or *dead space*, for which the motor does not run, and that the motor shall not over-run this dead space in stopping. If over-running does occur each time the motor operates, then the system will oscillate mechanically or *hunt*, usually over such a wide range as to be practically useless. Charging of smoothing capacitor C_0 through counter-load resistor r slows down response of the voltage at point A to changes in output of potentiometer r_2. The result is that, with any acceptably narrow dead space, overshooting and servo hunting would certainly occur without the damping circuit r_4, C_2.

It is well known from servo theory that hunting usually may be prevented by adding to the signal controlling the servo a component of proper polarity proportional to the rate of change of the servo output. The differentiating circuit r_4, C_2 of Fig. IV.-20 provides just such a component and is both simple and highly effective in permitting stable operation with a moderate dead space. The value of

C_2 required depends upon the inertia of the moving parts of the servo system and the stiffness of the servo-control action. This circuit provides the servo, in effect, with the ability to anticipate what its own action will do to the voltage at point A.

Dead space is damaging to the smoothness of servo action and may prove annoying or harmful. If it is made too narrow, however, there will be no stable potentiometer position for which the motor does not run and hunting will be inevitable. This difficulty can be overcome by connecting in series with C_2 a source of sinusoidal alternating voltage (for example a glow-tube oscillator) having a frequency which is higher than the natural hunting frequency of the servo but is still very low; the alternating voltage must have an amplitude E_a in excess of the range of relay-tube grid bias between the operating points of the two relays.

With this arrangement, when the voltage at A is balanced to zero, the relays will both operate at the frequency of the auxiliary alternating voltage E_a; the motor will run momentarily forward and momentarily backward on each cycle of that voltage. Forward and backward motions will be equal because the auxiliary-voltage variation will be symmetrical with respect to the operation of the two relays, and the average servo position will not change. Should the counter become unbalanced by too high a potentiometer output, however, the symmetry of relay operation will be disturbed, the high-current relay will be closed for a larger portion of the auxiliary-voltage cycle than the low-current relay will be open, and the motor will run longer in one direction than in the other. The average servo position will change in very small, rapid steps to rebalance the counter. Motor torque is not applied suddenly in full amount, but increases in average value in proportion to the error to be corrected. Forced servo vibration acts strongly to suppress the free hunting oscillation, but the damping circuit r_4, C_2 remains necessary because of the large back lash produced by time lag in the smoothing circuit r, C_0.

The vibrating servo has no dead space but nevertheless does have a true stable balance position for each value

of range. Large jumps in following are absent but the auxiliary-frequency wiggle of the servo shaft at balance can be annoying. This can be reduced as far as desired (within the limits of rapidity of relay action) by increasing the vibration frequency, but only at the expense of speed of smooth servo response. Because of the large alternating current through the motor even at balance, motor power must be considerably reduced in a vibrating servo to avoid overheating. Great sensitivity of balance is attainable by use of sufficient relay-amplifier gain.

Variation of internal potential (emission velocity, contact potentials, etc.) of the servo amplifier, V_5 of Fig. IV.-20, results in a wandering zero for the range scale of the servo. This is in contrast to the effect of internal-potential variations of limiter tube and counter diodes, which alter the effective limiter swing and thereby the scale factor of the counter but do not affect the scale zero. With a balanced amplifier circuit, zero variations can be reduced to one or two tenths of a volt. Counter output must be made sufficient to render the servo-zero wandering of relatively negligible importance.

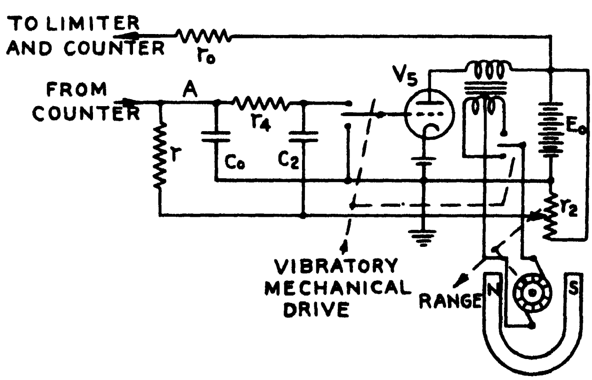


Fig. IV.-21. Smooth-acting range servo.

A very smooth, responsive servo, free of zero wander and motor overheating and practically as simple as that just described, is shown in Fig. IV.-21. It has the disadvantage of requiring considerable amplifier-output power, but the amplifier may of course have more stages than the single one shown. Departure of voltage at A from zero provides alternating amplifier input by way of a mechanical

inverter, and amplifier output is mechanically rectified in synchronism with input-inverter switching to provide direct motor-driving current. This current is zero at balance, and has a positive or negative average value respectively when there is a positive or negative unbalance voltage at *A*.

Balancing by servo control of counter sensitivities is possible in principle but might be inconvenient in practice and has not been used. Control of radar speed sensitivity would require adjustment of radio carrier frequency and is not easily practicable. Automatic control of radar range sensitivity by servo adjustment of frequency-modulation sweep width or of modulation frequency is entirely feasible and has been used experimentally.

Direct servo control of relative speed of radar and target has not been used, but control of range has been very useful and successful in automatic flight of aircraft at constant radar altitude. The very simple servo of Fig. IV.-20 has proved usable for this purpose, the radar servo motor acting to adjust the gyroscopic automatic pilot which in turn controls the elevator surfaces of the aircraft. Altitude to be maintained is selected by manual adjustment of the tap on voltage divider r_2; the servo makes slight changes in the position of this tap whenever it calls for control of the aircraft, simply to provide itself and the aircraft with anti-hunting signal. Because of the relatively sluggish response of an aircraft to its controls, there is a strong tendency to overshoot in altitude and the anticipatory action of the anti-hunt circuits is very important.

No difficulty is found in making a servo follow the output of a freely acting switched counter of the type shown in Fig. IV.-12. But if a switched counter operates over an appreciable output-voltage range, differential variations in diode bias of the upsweep and downsweep counters will upset to a troublesome degree the counter-sensitivity balance. Such operation is therefore to be avoided.

d. *Overall Result.* It is evident that a frequency-modulated radar system using switched counters is capable

of producing a total direct-voltage output

$$e = k_R h_R R - k_S h_S S + e_o \tag{IV.40}$$

dependent both upon range R and upon speed of closing S. Radar range sensitivity k_R, radar speed sensitivity k_S, range-counter sensitivity h_R, and negative speed-counter sensitivity h_S have the values given respectively by equations (II.22a), (II.23a), (IV.27) and (IV.28); e_o is bias voltage applied to the counter load. This output may be used to actuate a relay when it reaches a chosen value e_1, or to maintain itself at such a value by automatic servo adjustment of e_o, k_R, h_R, h_S, R, or S.

Accuracy of radar range sensitivity depends upon accurate control of rate of change of radio frequency produced by the modulator system, and that of speed sensitivity upon mean radio frequency. These sensitivities are both scale factors; errors in them correspond to definite fractional errors in range and speed, but do not shift the zeros from which range and speed are measured. The radar system of course measures total range traversed by its signals, including the electrical length of any radio-frequency cables used to connect transmitter and receiver to their respective antennas; such cables do affect the zero of the range scale and must be taken into account in the use of the equipment.

Counter range and speed sensitivities also directly affect the accuracy of the system, again as scale factors. The counter circuits, like the modulator and its circuits, must therefore be carefully designed and built, using reliable components, if good overall accuracy is to be maintained. Bias voltage e_o and relay-threshold or servo-balance voltage e_1 can introduce errors of the zero-shift type although they do not appear as scale factors; accuracy of these voltages is also required.

In pulse radars arranged for precision ranging, errors in range tend to be constant in amount, independently of range. In f-m radar, with important potential sources of scale-factor error, range errors are likely to correspond to a constant fraction of the range. This suggests that f-m radar is likely to be the more useful at short ranges. Range measurement accurate to a few per cent by means of

f-m radar is practicable without extensive precautions. The servo-balanced null type of counter, without switching, is by itself capable of fractional-per-cent accuracy. Where speed frequency exceeds range frequency, fractional-per-cent overall accuracy in determination of speed alone is feasible. Where range frequency exceeds speed frequency, high accuracy in determination of range alone is possible by the use of accurate modulation. Accuracy practically attainable in combined range and speed determination by use of switched counters has never been clearly determined but is probably a few per cent.

6. NOTATION AND REFERENCES

a. *Notation.* The algebraic and circuit notation used in this chapter is listed alphabetically below.

c	Velocity of radio-wave propagation.
C	Charge-transfer capacitance of averaging cycle-rate counter.
C_0, C_1	Counter-output smoothing (integrating) capacitance.
C_2	Servo-damping capacitance.
C_n, C_p	Charge-transfer capacitance of negative-output and positive-output switched counters respectively.
C_R, C_S	Charge-transfer capacitance of range and speed counters respectively.
e	Counter-output voltage.
e_0	Bias voltage added to counter output.
e_1	Special value of counter output, including bias, to cause some operation to take place.
e_d	Voltage difference between charge and discharge terminals of counter tubes.
e_i	Internal voltage of counter tube (electron-emission velocity, contact potentials, etc.)
E_0	Supply voltage to counter.
E'_0	Effective supply voltage, allowing for E_1, e_d and e_i.
E_1	Minimum voltage of limiter plate.
E_a	Amplitude of alternating voltage used to force servo vibration.
f	Frequency of signal to counter.
f_0	Reference value of varying frequency.
f_m	Modulation frequency of f-m radar.
f_R, f_S	Range and speed beat frequencies of f-m radar.

$\dot{f}$	Time rate of change of varying frequency.
F_o	Carrier frequency of radio signal.
h_R	Range sensitivity of counter in volts per cycle per second.
h_S	Speed sensitivity of counter in volts per cycle per second.
$h''_R, h''_S,$ h'_S	Counter range and speed sensitivities as modified by characteristics of counter switching.
i	Counter-output current.
i_d, i_u	Counter current during downsweep and upsweep respectively of frequency modulation.
I	Maximum plate current of pentode limiter that is not affected by grid voltage in negative region.
k_R	Radar range sensitivity in cycles per second per unit range.
k_S	Radar speed sensitivity in cycles per second per unit speed.
M	Current-indicating meter.
N_o	Number of standing waves between radar and target for radio frequency F_o.
N_W	Number of standing waves between radar and target if radio frequency were W.
ΔN	Change in number of standing waves as radio frequency is modulated from $F_o - \frac{1}{2} W$ to $F_o + \frac{1}{2} W$.
q	Charge transferred by counter capacitor.
q_n, q_p	Charge transferred by capacitors of negative-output and positive-output counters respectively.
r	Counter load resistor.
r_o	Limiter plate resistor.
r_1	Cathode-follower cathode resistor.
r_2	Null-counter balancing potentiometer.
r_3	Relay-bias smoothing resistor.
r_4	Servo-damping resistor.
$r_a, r_b,$ r_p, r_s	Resistors in auxiliary circuits.
R	Range or distance between radar and target.
S	Relative approach speed of radar and target.
t	Time
t_o	Reference instant of time.
V_1	Vacuum tube limiter.
V_2	Charging tube of counter.

V_3 Discharging tube of counter.

V_4 Cathode follower tube.

V_5 Relay-actuating tube.

W Width of band swept in frequency modulation.

γ Fraction of complete switching cycle that both counters of switched system are turned off on each half of switching cycle, or blanking fraction.

δ Fraction of switching cycle by which modulation turn around lags counter switching.

ϵ Fraction of switching cycle by which on time of one counter exceeds and on time of other counter falls short of equal duration.

λ_o Radio-signal wave length for frequency F_o.

λ_W Radio-signal wave length for frequency W, or sweep wave length.

$\Delta\psi$ Change in phase of received signal relative to mixing signal during modulation.

b. *References.*

1. P. Giroud and L. Couillard: "Sondeur Radioelectrique pour la Mesure des Hauteurs des Aeronefs au-dessus du Sol," *Ann. de Rad. Elect.*, Vol. II., no. 8, pp. 150-172 (April 1947).

2. W. C. Wilkinson: "Investigation of a Discriminator Type of Frequency Indicator," report no. 36 under contract NXsa-35042 (written Sept. 1944, issued Jan. 15, 1947).

3. H. Lauer, R. Lesnick and L. E. Matson: *Servomechanism Fundamentals*. New York, McGraw-Hill, 1947.

CHAPTER V.

MOTION OF AIRCRAFT AND MISSILES

1. General Considerations

In order to understand the design of frequency-modulated radar systems for bombing and other special uses, it is necessary first to become familiar with some aspects of the motions of aircraft and missiles at low altitudes. As little as possible will be said here of the forces controlling the motions, that is, of the dynamics of the situation. The main consideration will be certain properties of the motions themselves, that is, of the kinematics peculiar to f-m radar applications. Much of the discussion will be based on certain time intervals which are conveniently and directly related to the data provided by f-m radar and to the control of bomb release.

2. Direct Approach

a. *Approach Kinematics.* The simplest case to consider is that in which the radar is moving at constant speed directly toward its target. Time interval T which must elapse between the instant at which a radar observation is made and the instant at which the radar strikes the target is given directly, in this simple case, by the ratio of the observed range or distance R between radar and target to the observed speed S with which the two are approaching. For any given time-to-target T, range and speed are related by the equation

$$R/T - S = 0 . \qquad \text{(V.1)}$$

Comparison of equation (V.1) with overall f-m radar action as described by equation (IV.40) shows such similarity that the suitability of f-m radar for providing information concerning this particular motion is immediately evident. If, for example, a servo controlling the range sensitivity of the radar maintains the total counter-output voltage e equal to the counter bias e_0, then a continuous determination of the time required to reach the

target will be made automatically. The time to target will then be given at each instant by the ratio of overall speed and range sensitivities of the radar system. Thus

$$T = k_S h_S / k_R h_R , \tag{V.2}$$

where k_S, h_S and h_R are fixed at predetermined values and k_R is set by the servo.

b. ***Bombing from Direct Approach.*** A highly artificial case will serve as a very simple introduction to some factors found in actual bombing approaches. Let a radar target at B of Fig. V.-1 be suspended (for example from a tower or a balloon) directly above a surface target at Q, which is to be bombed by an aircraft at P in level flight directly toward and at the same altitude as the radar target. Neglecting air resistance, no horizontal force

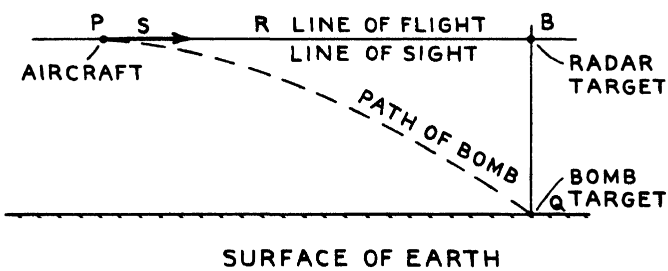


Fig. V.-1. Geometry of direct approach.

will act on the bomb after its release from the aircraft; the bomb will therefore continue to travel forward at uniform speed, as will the aircraft, and so will remain directly under the latter as it falls.

The interval of time T_f taken by a bomb to fall from altitude A is simply

$$T_f = \sqrt{2A/g} , \tag{V.3}$$

where g is the constant downward acceleration produced by gravity. If, therefore, a bomb is released from the aircraft at an instant that precedes the collision of aircraft and radar target by a time interval which is exactly the time of fall T_f, it will fall directly below the aircraft and will strike the surface and the surface target just when the aircraft strikes the radar target at its own altitude.

By setting the range sensitivity of an f-m radar on the bombing aircraft to the value

$$k_R = k_S h_S / (h_R \sqrt{2A/g}), \tag{V.4}$$

and arranging for relay actuation when total output e of a switched counter operated by the radar beat note falls to the counter-bias value e_0, bombs may be released automatically at the correct instant to strike the surface target. Release range will thus be so determined, in accordance with equation (V.1), that the instant of release will remain correct whatever the speed S of closing of the aircraft on its target. This state of affairs will not be upset by wind or target motion, since it depends only upon relative speed of aircraft and target and this as well as range is measured directly by the radar.

The relation between speed and range at release is shown for several values of aircraft altitude (or time of fall of bomb) in Fig. V.-2. It is required that the aircraft be

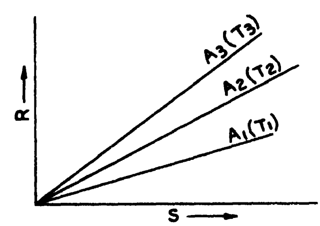


Fig. V.-2. Range-speed relation for direct approach.

flown so as to approach directly the elevated target at any constant speed, while flying level at the altitude for which the radar sensitivity is set. The interest of this artificial special case lies in the simplicity of these speed-range relationships and in the direct applicability of f-m radar data to provide automatic bomb release at exactly the correct moment for a hit.

3. LEVEL FLIGHT APPROACH

a. ***Approach Kinematics.*** A more usual problem than that of direct level approach terminating in collision is that of an aircraft flying level in such a direction as to pass directly over a surface target. The geometry of this

case is illustrated by Fig. V.-3, where an airplane at point P is at an altitude A above the land or sea surface and at slant range R from a surface target at point Q. The

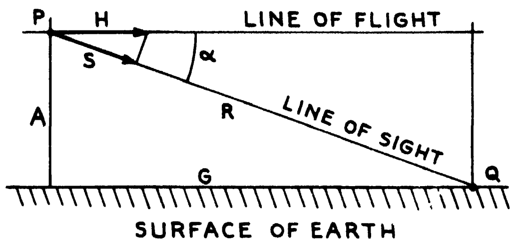


Fig. V.-3. Geometry of level-flight approach.

projection of R on the ground is the ground range G separating aircraft and target. Slant closing speed S of aircraft relative to target is measured along the line of sight, while the horizontal closing speed H is measured along the line of flight.

Closing speed H of aircraft relative to vertical through target is the algebraic sum of the components in the vertical plane through aircraft and target of target velocity over the earth, wind velocity over the earth, and air speed (or rather velocity) of the aircraft; all of these component velocities are horizontal in the case of level flight. Any relative-velocity component not in the vertical aircraft-target plane must be avoided; this is a matter of aim rather than of ranging and is not of interest at the moment.

With flight maintained at constant horizontal speed H, as is normal in level approach, an aircraft at ground range G will pass over the target after a time interval T given by

$$HT = G. \qquad (V.5)$$

The radar, however, does not measure ground range G and ground speed H but rather slant range R and slant speed S. Comparison of the distance- and velocity-vector right triangles in the vertical plane represented by Fig. V.-3 shows them to be similar, so that

$$H/S = R/G, \qquad (V.6)$$

while, of course,

$$R^2 = G^2 + A^2. \qquad (V.7)$$

Eliminating G and H from equation (V.5) by use of (V.6) and (V.7), there results

$$R/T - S = A^2/(RT)\,. \tag{V.8}$$

This represents a considerably more complicated state of affairs than the direct approach of (V.1).

Dependence of slant range on slant speed, as given by (V.8), is shown for two arbitrarily chosen values of altitude A and two arbitrary values of time-to-target-crossing T by the curves of Fig. V.-4(a). As would be expected,

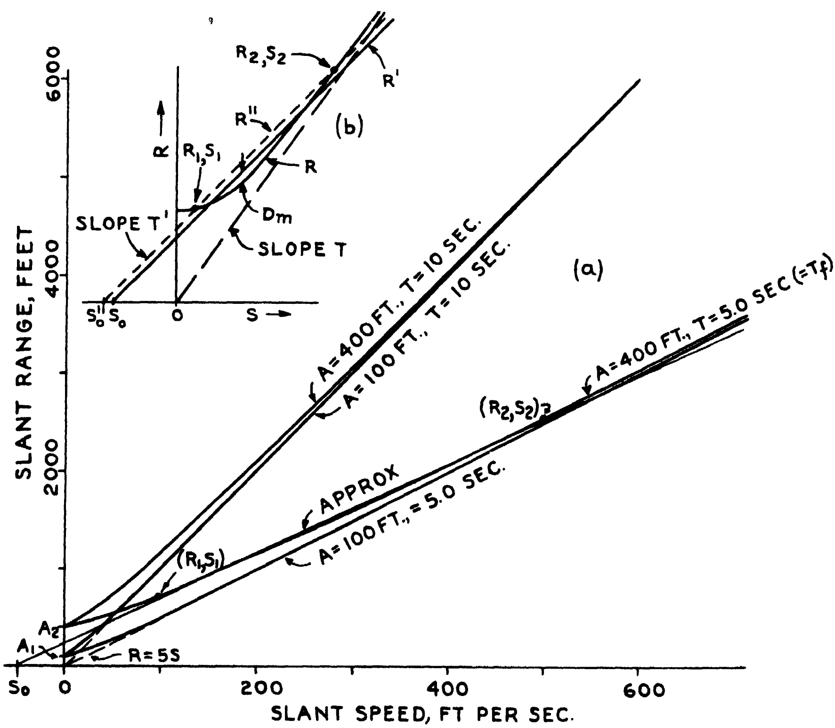


Fig. V.-4. Slant range-slant speed relations for level flight.

the relationship always approaches the simple linear one of slope T, represented by Fig. V.-2 or the dashed lines of Fig. V.-4, for slant ranges much greater than the approach altitude. For the lower altitude illustrated, the curves become indistinguishable from these lines even at quite short range. For vanishing slant speed, which corresponds

to a position of the aircraft directly over the target, the range does not fall to zero but only to the value of the altitude A. Such a range-speed relationship presents a problem not exactly soluble by the radar-signal utilization means described in Chapter IV.

Examination of Fig. V.-4(a) shows, however, that the range-speed curve for a given altitude and time to target may be approximated very closely indeed, over a considerable region of speed variation, by a straight line. This is true so long as excessively low speeds and high altitudes are avoided, and permits use of the radar system to provide automatically a useful approximate solution. The slope and intercept of each approximating line, such as the light full line of the figure, is determined by the limits H_1 and H_2 of horizontal closing speed between which best approximation is desired, as well as by the values of altitude and time to target for which the slant range-slant speed relation is to be approximated. Radar range sensitivity and counter bias voltage to be used are in turn controlled by these required slope and intercept values for the approximating line.

The limiting horizontal closing speeds between which approximation is required are determined by the ranges of aircraft air speeds, speeds of target motion, and wind speeds over which it is desired to operate. Minimum horizontal closing speed H_1 is the minimum operating air speed of the slowest aircraft considered, minus the speed of the maximum head wind under which operation is required, minus the maximum speed at which a surface target is expected to move away from the attacking aircraft. Maximum horizontal closing speed H_2 is the maximum operating air speed of the fastest aircraft considered, plus the speed of the maximum tail wind anticipated, plus the maximum speed at which the target may move toward the aircraft. Once determined by these considerations of limits of operation, H_1 and H_2 become basic design constants.

Given lower and upper closing-speed limits H_1 and H_2, dependence of slope and intercept of an approximating line on altitude and time to target must be determined. This is most conveniently done in terms of the angle α by which the line of sight (and of radar transmission) from aircraft

to target is depressed below the horizon, given simply by

$$\tan\alpha = A/G = A/(HT). \tag{V.9}$$

Limiting values α_1 and α_2 are determined respectively by $A/(H_1T)$ and $A/(H_2T)$; it should be noted that α_1 and α_2 do not depend on altitude and time to target separately, but only on the ratio A/T. Referring again to the distance and speed triangles of Fig. V.-3, it should further be noted that

$$R = A/\sin\alpha \tag{V.10}$$

$$S = H/\cos\alpha\ . \tag{V.11}$$

By analogy with the case of direct approach, for which the slope of each range-speed line of Fig. V.-2 is the corresponding time to target T, the slope of an approximate range-speed line in Fig. V.-4 may be called T'. To distinguish it from the true range R given by a curve of Fig. V.-5 [equation (V.8)], the approximate range given by a corresponding line may be called R', as indicated in exaggerated fashion in Fig. V.-4(b). The intercept of an approximation line on the speed axis, which can be seen from Fig. V.-4 to have a negative value, may be called S_0. The equation for the approximate range is then

$$R'/_{T'} - (S-S_0) = 0\ . \tag{V.12}$$

For the limiting horizontal approach speeds H_1 and H_2, the corresponding values R_1 and R_2 of slant range and S_1 and S_2 of slant speed may be found from equations (V.9), (V.10), and (V.11) for any chosen value of A/T. The average slope of the range-speed curve between points (R_1, S_1) and (R_2, S_2) is, by definition,

$$T' = (R_2 - R_1)/(S_2 - S_1). \tag{V.13}$$

Expressing R_1, R_2, S_1, and S_2 of this equation in terms of α_1 and α_2, and simplifying the complex fraction which results, T' is found to be simply

$$T' = T/(1 + \sin\alpha_1 \sin\alpha_2)\ . \tag{V.14}$$

Designating the range given by the straight line through (R_1, S_1) and (R_2, S_2) as R'' and the intercept of this line as S_0'', as in Fig. V.-4(b),

$$R'' = (S - S_0'')T'. \tag{V.15}$$

Since R_1'' (or R_2'') is the same as R_1 (or R_2), the values of R and S in terms of α for either limiting speed may be used in (V.15) to find that

$$S_o'' = -(A/T)(\sin\alpha_1 + \sin\alpha_2)\ . \qquad \text{(V.16)}$$

From equations (V.10), (V.11), (V.14), (V.15), and (V.16), the departure of R'' from R, as indicated in Fig. V.-4(b), is

$$2D = R'' - R = A\,\frac{\sin\alpha_1 + \sin\alpha_2 - \sin\alpha - (\sin\alpha_1\sin\alpha_2/\sin\alpha)}{1 + \sin\alpha_1\sin\alpha_2}\ . \qquad \text{(V.17)}$$

This departure is, of course, zero for line-of-sight depression angle α_1 or α_2, as a result of the definition of R''. As may be verified by equating to zero the derivative $dD/d\alpha$, D has the maximum value

$$D_m = \tfrac{1}{2}A(\sqrt{\sin\alpha_1} - \sqrt{\sin\alpha_2})^2/(1 + \sin\alpha_1\sin\alpha_2) \qquad \text{(V.18)}$$

at that depression angle for which

$$\sin\alpha = \sqrt{\sin\alpha_1\sin\alpha_2}\ . \qquad \text{(V.19)}$$

Examination of Fig. V.-4(b) will show that the maximum error of approximation is cut in half by using, instead of the line R'', a line giving an approximate range R'. This line has the same slope T' as that giving R'', but is moved downward by D_m, so that

$$R' = R'' - D_m \qquad \text{(V.20)}$$

and

$$S_o = S_o'' + D_m/T', \qquad \text{(V.21)}$$

or

$$S_o = -\tfrac{1}{2}(A/T)(\sqrt{\sin\alpha_1} + \sqrt{\sin\alpha_2})^2. \qquad \text{(V.22)}$$

At the ends of the operating speed range, points (R_1,S_1) and (R_2,S_2) of Fig. V.-4, the error $R'-R$ due to the linear approximation is $-D_m$, while the error is $+D_m$ at that mid-range speed for which (V.19) is satisfied. As may be seen from Fig. V.-4(a), which is drawn to scale, these maximum errors are small. For all other speeds in the chosen operating range, the errors resulting from use of the linear approximation are still smaller.

Both the slope-correction factor T/T' and the speed intercept S_o of the linear range-speed approximation have

now been determined [equations (V.14) and (V.22)]. Each depends only on the parameter A/T, for any specific upper and lower limits of horizontal approach speed [see equation (V.9)]. Both S_0 and A/T, being speeds, may be specified by their ratios to the geometric mean $\sqrt{H_1H_2}$ of the operating approach-speed limits. The maximum range error due to the approximation may likewise be expressed as a fraction of the ground range $\sqrt{H_1H_2T}$, or $\sqrt{G_1G_2}$. This is the ground range of a position requiring the lapse of time T for the aircraft at mid speed to reach a point directly over the target. Fig. V.-5(a), (b) and (c) show respectively slope

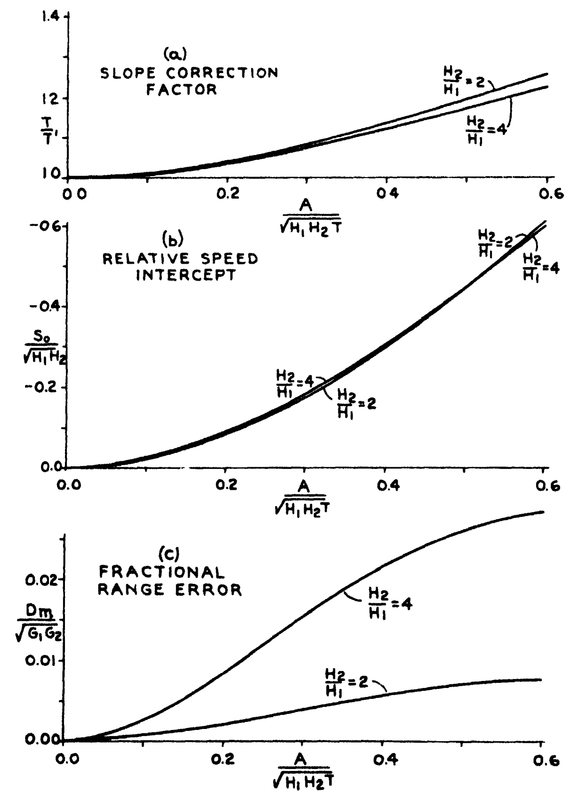


Fig. V.-5. Characteristics of linear approximation.

correction T/T', relative speed intercept $S_0/\sqrt{H_1H_2}$, and maximum fractional range error $D_m/\sqrt{G_1G_2}$ as functions of $A/(\sqrt{H_1H_2}T)$, which is $\sqrt{\tan\alpha_1\tan\alpha_2}$. Each of these curves is fully determined by and marked with a chosen ratio H_2/H_1

of operating speed limits. Methods of producing physically the quantities T/T' and S_o as functions of A/T will be described in a later chapter on f-m radar systems.

b. ***Bombing from Level Approach.*** In bombing from level flight it is again only necessary, so long as moderate speed and low altitude permit neglect of windage effects, to release the bomb at that instant which precedes the passage of aircraft over target by the time of fall of the bomb. For bombing, therefore, the time to target to be used in determining a_1 and a_2, as well as in the remainder of equations (V.14) and (V.22), is the time of fall given by equation (V.3). The only variable parameter then determining the coefficients of equation (V.12) and correspondingly the settings of the radar system is the altitude of flight. This is the case for the 400-foot, 5-second curve of Fig. V.-4, since the time of fall from an altitude of 400 feet is 5 seconds.

Two corrections are required by radar characteristics, however; these have been neglected thus far to simplify the initial discussion. As pointed out in section 2e of Chapter IV., there is a time lag due to smoothing in the measurement of range and speed by f-m radar. Allowance must therefore be made for a time lag τ_s caused by smoothing of counter output and smoothing of relay input, as well as by action of relays and action of bomb-release mechanism. It was also pointed out earlier that the radar measures not only true target range but, in addition, the electrical length of its own radio-frequency cables. Allowance must also be made for this *residual range* R_r.

To allow for time lag, it is only necessary to set up the radar to release the bomb at a range which corresponds, for the measured slant speed, to a time to target of $T_f + \tau_s$. Arrival of the aircraft at this range, $T_f + \tau_s$ seconds before it will cross the target, then initiates a sequence of counting, smoothing, relay and release operations which will cause the bomb actually to start its fall just τ_s seconds later. Falling for the remaining T_f seconds, the bomb then strikes the surface — and the target — just as the aircraft passes over the target. The slope T' of the linear approximation on which the radar works, related to the radar sensitivity factors as in

equation (V.2), should therefore be made

$$T' = (\sqrt{2A/g} + \tau_s)/(1 + \sin\alpha_1 \sin\alpha_2), \tag{V.23}$$

where α_1 and α_2 are given by

$$\left.\begin{aligned} \tan\alpha_1 &= A/[H_1(\sqrt{2A/g} + \tau_s)] \\ \tan\alpha_2 &= A/[H_2(\sqrt{2A/g} + \tau_s)]. \end{aligned}\right\} \tag{V.24}$$

Instead of the slant range R of Fig. V.-3, the radar measures $R + R_r$. If the total length of all radio-frequency cables is L and the radio-wave velocity on these cables is v, then the corresponding wave-travel time delay is L/v. The travel-time delay for two-way space transmission over range R_r is $2R_r/c$. The residual range equivalent to the cables is therefore

$$R_r = cL/(2v). \tag{V.25}$$

For the release-producing sequence to be initiated at the approximated range R', with radar actually measuring a range $R' + R_r$, the release condition becomes

$$(R' + R_r)/T' - S = R_r/T' - S_0. \tag{V.26}$$

The kinematic condition (V.26) for bomb release may be compared with the electrical range-speed relation of equation (IV.40), for which an f-m radar system will actuate a relay. In terms of counter-bias voltage e_0, total counter output e_1 to initiate release, and speed-sensitivity factors k_s and h_s, the radar may be used to release a bomb if $e_1 - e_0$ is so adjusted that

$$R_r/T' - S_0 = (e_1 - e_0)/(k_s h_s). \tag{V.27}$$

H_1 and H_2 are constants of specified operation and τ_s, R_r, k_s and h_s are constants of the radar installation, while S_0 for bombing is

$$S_0 = -\tfrac{1}{2}A\,(\sqrt{\sin\alpha_1} + \sqrt{\sin\alpha_2})^2 / (\sqrt{2A/g} + \tau_s) \tag{V.28}$$

and T' is still as given by (V.23).

For an operating speed range of 175 to 700 feet per second (approximately 100 to 400 knots) and a delay τ_s of 0.40 second, the value of $A/(\sqrt{H_1 H_2}\,T)$ at an altitude of 400 feet is 0.212. Fig. V.-5(c) indicates under these

conditions a fractional range error of 0.91 per cent, or an actual range error of 16 feet at the mid-speed ground range for release of 1750 feet. In view of the many error sources active in fully automatic bombing of invisible targets, the above range error contributed by use of a linear range-speed approximation seems quite acceptable. A rudimentary analysis to determine bombing errors resulting from slight errors in determination of R, S and A, or in setting S_0 and T', has indicated that the type of data considered in this chapter is markedly satisfactory for control of bomb release. More complete error analysis is needed, to permit a real comparison of accuracy to be made between this and other methods of controlling release of bombs.

4. VERTICAL MANEUVERING

a. *Kinematics.* An aircraft approaching a target in a steady climb or dive moves as indicated in Fig. V.-6, which differs from Fig. V.-3 only in including a vertical speed component V, considered positive when directed upward as in climbing. So long as the horizontal closing-speed component H does not change during approach, the time to elapse before the target is crossed is still related to present ground range G by the simple equation (V.5).

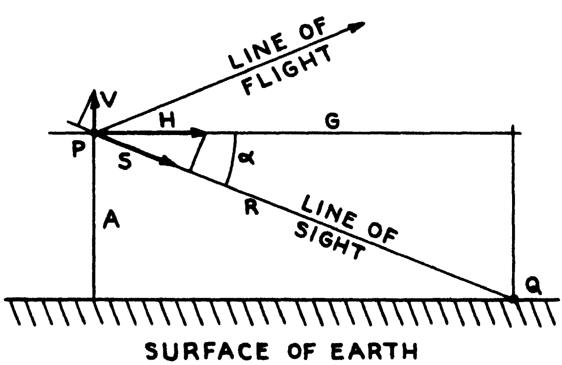


Fig. V.-6. Geometry of approach with vertical speed.

From the similar speed- and range-vector triangles of Fig. V.-6,

$$H = \left(S + {}^{VA}/_{R}\right){}^{R}/_{G}. \tag{V.29}$$

Therefore, using (V.5) and (V.7),

$$R/T - S = \left(A/T + V\right) A/R\,, \tag{V.30}$$

rather than the simpler equation (V.8), relates slant range and slant speed for any chosen values of altitude, vertical speed, and time to target. The graph of equation (V.30) is similar to that of (V.8), except that the minimum range for zero slant speed is no longer A but $A\sqrt{1 + VT/A}$. Fig. V.-4 therefore remains qualitatively applicable, and linear approximation giving a range R' remains permissible.

Equations (V.9) for depression angle α and (V.10) for R remain valid, but (V.11) must be replaced by

$$S = H\cos\alpha - V\sin\alpha\,. \tag{V.31}$$

With the use of (V.31) instead of (V.11), the same procedures used for the case of level flight serve again to determine slopes and intercepts of the approximating lines, as well as maximum difference between approximate and true ranges. This time,

$$T' = T/\left\{1 + \left(1 + VT/A\right)\sin\alpha_1\,\sin\alpha_2\right\}\,, \tag{V.32}$$

$$S_0'' = -\left(A/T + V\right)(\sin\alpha_1 + \sin\alpha_2)\,, \tag{V.33}$$

$$D_m = \tfrac{1}{2}(A + VT)(\sqrt{\sin\alpha_1} - \sqrt{\sin\alpha_2})^2/\left\{1 + \left(1 + VT/A\right)\sin\alpha_1\sin\alpha_2\right\} \tag{V.34}$$

and

$$S_0 = -\tfrac{1}{2}\left(A/T + V\right)(\sqrt{\sin\alpha_1} + \sqrt{\sin\alpha_2})^2\,; \tag{V.35}$$

the depression angle for maximum error is again given by equation (V.19). These expressions obviously revert for zero vertical speed to (V.14), (V.16), (V.18) and (V.22), respectively. Each curve of Fig. V.-5 is simply the zero-vertical-speed member of a family of similar curves, each family representing one of the equations (V.32), (V.35) and (V.34) for various constant values of vertical speed.

b. ***Bombing.*** Bombing from an inclined flight path, called dive or toss bombing according to the downward or upward slope of the path, is often decidedly advantageous.

Time of fall of the bomb in such a case, again neglecting air resistance, is

$$T = \sqrt{2A/_{g} + V^2/_{g^2}} + V/_{g} \; , \qquad \text{(V.36)}$$

where vertical speed V is positive for climbing flight. The bomb again remains directly below the aircraft that released it, so long as the latter does not change the horizontal component of its speed after release. Release must therefore again occur prior to passage of aircraft over target by an interval equal to the time of fall of the bomb.

The value of altitude used to determine the geometric corrections due to depression of the line of sight should be that existing at the instant of measuring slant range and slant speed. Time of fall, on the other hand, must be determined from the aircraft altitude at the instant of release. Let τ_s again represent time lag between the instant of observation of range and speed and the instant of release, but now let τ_a represent in addition the corresponding instrumental time lag from the instant of occurrence of a particular altitude until this altitude datum is made available for use. The sequence of events in time is then as shown in Fig. V.-7. All data must be available at the instant, τ_s seconds before release, when

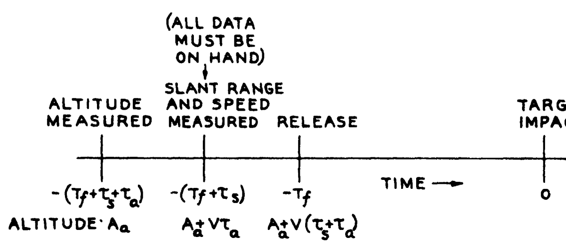


Fig. V.-7. Sequence of events in bombing.

the final slant-range datum is accepted by the radar. Range sensitivity and counter bias must be already finally set at this instant. Counter smoothing will introduce the same time lag in output response to bias changes as it does to counter-current changes.

Considering vertical speed not to change in the small time intervals involved, and letting A_a be the aircraft altitude at the instant of final altitude observation, the

altitude when range is measured will be $A_a + V\tau_a$ and that at release will be $A_a + V(\tau_a + \tau_s)$. Range data should initiate release when range reaches the value corresponding to a time to target of $T_f + \tau_s$, given by

$$T_f + \tau_s = \sqrt{2A_a/g + 2V(\tau_a + \tau_s)/g + V^2/g^2} + V/g + \tau_s. \qquad (V.37)$$

Inserting proper values of A and T in equation (V.32), the slope of the linear range-speed approximation is

$$T' = (T_f + \tau_s)\Big/\left\{1 + \left[1 + V(T_f + \tau_s)/(A_a + V\tau_a)\right]\sin\alpha_1 \sin\alpha_2\right\}, \qquad (V.38)$$

where

$$\left.\begin{aligned} \tan\alpha_1 &= (A_a + V\tau_a)/[H_1(T_f + \tau_s)] \\ \tan\alpha_2 &= (A_a + V\tau_a)/[H_2(T_f + \tau_s)]\,. \end{aligned}\right\} \qquad (V.39)$$

Similar insertion of values into (V.35) determines speed intercept S_0 as

$$S_0 = -\tfrac{1}{2}\left[(A_a + V\tau_a)/(T_f + \tau_s) + V\right](\sqrt{\sin\alpha_1} + \sqrt{\sin\alpha_2})^2. \qquad (V.40)$$

Equations (V.38) and (V.40) would be quite complicated if written out fully in terms of A_a, V, H_1, H_2, τ_a, and τ_s; they signify simply enough that the correct slant-range-slant-speed relation for release is fully determined, even for toss or dive bombing, by measured aircraft altitude and vertical speed (given the apparatus constants τ_a, τ_s, H_1, and H_2). Allowance to be made for residual range due to cable length is the same as in the case of level approach.

c. ***Effect of Acceleration.*** To allow the freedom of vertical maneuvering necessary for an evasive approach, vertical acceleration of the bombing aircraft is necessary. Such acceleration has in general no effect on the actual release or fall of the bomb, because the bomb becomes completely free of the aircraft upon release and thereby ceases immediately to be affected by acceleration of the aircraft. Acceleration does act to alter the vertical speed of the aircraft during the time lag τ_s, as well as during the time lag τ_v from occurrence of a particular

vertical speed V_v to availability of this speed datum, and must therefore be taken into account. Altitude is also altered during time lags τ_a and τ_s, but usually to only a slight extent.

Vertical acceleration will be considered not to change significantly in the short intervals involved. Horizontal acceleration, changing the magnitude of H, is not essential to evasive action and must be avoided. Vertical acceleration a (positive upward) will change the data sequence of Fig. V.-7 to that of Fig. V.-8, where V_v is the value of vertical speed at the instant it is last usefully observed.

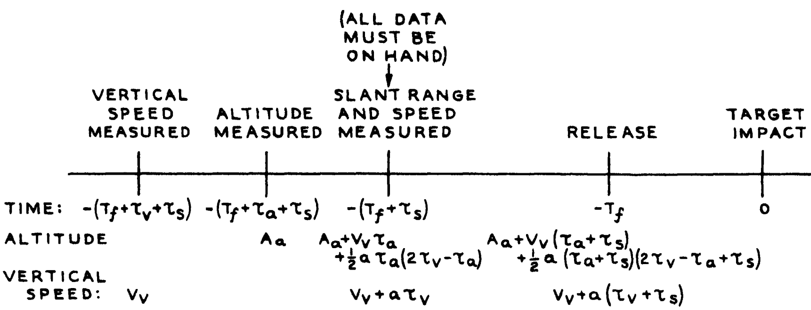


Fig. V.-8. Sequence of events in bombing, with vertical acceleration.

Time of fall still is given by equation (V.36), using the values of altitude and vertical speed that will occur at the actual moment of release, which are

$$\left.\begin{aligned} A_r &= A_a+V_v(\tau_a+\tau_s)+\tfrac{1}{2}a(\tau_a+\tau_s)(2\tau_v-\tau_a+\tau_s) \\ V_r &= V_v+a(\tau_v+\tau_s) \end{aligned}\right\} . \qquad (V.41)$$

The value of T to be used in (V.32) and (V.35) for determining T' and S_0, and in (V.9) for a, is $T_f+\tau_s$. Values of altitude and vertical speed to be used in (V.32) and (V.35) are those that will occur at the moment of making the final observation of slant range and speed, which are

$$\left.\begin{aligned} A_s &= A_a+V_v\tau_a+\tfrac{1}{2}a\tau_a(2\tau_v-\tau_a) \\ V_s &= V_v+a\tau_v \end{aligned}\right\} . \qquad (V.42)$$

Unless vertical accelerations are maintained long enough to result in considerable changes of altitude, no effective

evasive action results. That is, rapid and frequent reversals of flight-control effort defeat their own intent and will normally be avoided. Flight paths with altitude varying roughly sinusoidally, with a rather long period and a large amplitude, are probable and useful. This supports the above assumption that vertical acceleration will not change significantly during the short time lags of radar operation.

For a bomb released during a steep climb, the time of fall is substantially increased by the vertical speed, so release must occur early or at long range. Likewise, a release during a dive must be made at short range. If the correct release was not reached in a given climbing section of the approach, the marked decrease of range for release on going into the following dive makes it unlikely that the release point will be reached during that dive. Rapid increase of correct range for release during the pull-up following the dive makes a release in that phase of the approach probable. In a vertically wavy approach, therefore, it is improbable that release will occur during the diving portions.

Downward acceleration of the aircraft has some interesting properties. The effect of downward acceleration for which allowance must be made is, as in the case of upward acceleration, merely to alter vertical speed V, altitude A, and through them time of fall T_f, between the instants at which particular values of A and V occur and the earliest instant at which the radar equipment can cause bomb release based on those particular values. But if the downward acceleration exceeds that of gravity, the aircraft with bomb attached is falling faster than would a free bomb. If a bomb were released under such a condition, it could not fall from but would rather be pushed down by the aircraft.

Time of fall decreases rapidly under strong downward acceleration, whatever the altitude or vertical speed. During one second of flight with downward acceleration exceeding that of gravity, the aircraft comes only one second closer to passing over the target. But during the same second of flight, the time required for a bomb released from the aircraft to fall to the ground decreases by more

than one second. The proper moment for release is therefore receding into the future throughout the duration of any maneuver in which a downward acceleration exceeding one *g* is maintained. If release has not occurred before such a maneuver is begun, it cannot occur until after the maneuver is completed and *a* has returned to an algebraic value greater than $-1g$.

High downward acceleration before release inhibits the release while it is applied, but does not in any way disturb the occurrence of a normal release after the downward acceleration is reduced. High downward acceleration just after release, however, might cause the aircraft to overtake the falling bomb and disturb its motion. High upward acceleration does not affect the possibility of release, but does require strong correction of observed *A* and *V* for time lag in equipment.

d. *A Simpler Approximation.* Change in time of fall of the bomb is mainly responsible for the effect of vertical speed on release range. Fig. V.-9, depicting equation

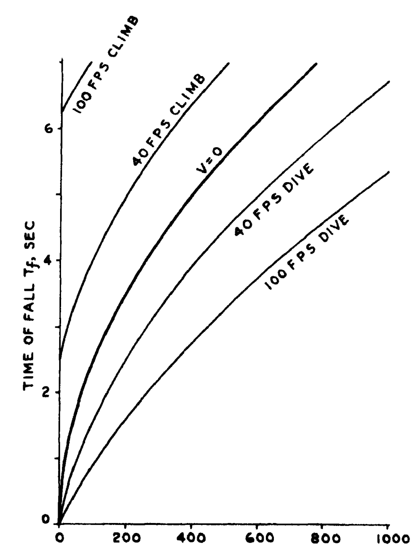


Fig. V.-9. Effect of vertical speed on time of free fall.

(V.36), shows how strongly vertical speed affects time of fall. Corrections for depression of line of sight are

small anyway, and slight errors in such corrections as a result of moderate vertical speed do little damage. The vertical speed required to upset time of fall in the approximation developed in section 3 for level approach is small enough to be hard to avoid. Fortunately, a rather simple approximate correction is possible for such small vertical aircraft speeds as may occur unintentionally.

Vertical speed developed in free fall through altitude A starting from rest is

$$V_0 = \sqrt{2Ag}\ \text{(downward)}. \tag{V.43}$$

A body starting from rest will fall from some altitude A' in the same length of time as is required to fall from A when starting with upward vertical speed V. Equating the time of fall given by equation (V.3) for A' to that given by (V.36) for A and V, and using (V.43),

$$A' = A\left\{1 + 2\,{}^{V^2}\!/_{V_0^2} + 2\,{}^{V}\!/_{V_0}\sqrt{1 + {}^{V^2}\!/_{V_0}}\right\} \quad . \tag{V.44}$$

For small vertical speeds, $V \ll V_0$, A' is approximately

$$A' = A\left\{1 + 2\,{}^{V}\!/_{V_0}\right\} \quad . \tag{V.44a}$$

Remembering that V is dA/dt, and taking into account (V.43), (V.44a) becomes

$$A' = A\left\{1 + 2\sqrt{{}^{2}\!/_{g}} \cdot {}^{d\sqrt{A}}\!/_{dt}\right\} \tag{V.45}$$

or, to the same degree of approximation,

$$A' = A\left\{1 - 2\sqrt{{}^{2}\!/_{g}} \cdot {}^{d\sqrt{A}}\!/_{dt}\right\}^{-1} \quad . \tag{V.45a}$$

This value of A' used in (V.3) determines the actual time of fall to a good approximation for such small vertical speeds as may occur unintentionally during "level" flight. In certain radar altimeters, a voltage quite closely proportional to $\sqrt{A}$ exists, and the above expression suggests the possibility of compensating very simply for changes in time of fall caused by small vertical speeds.

The coefficient ${}^{A}\!/_{T} + V$ determining speed intercept S_0 [equation (V.35)] is at release, to the same degree of

approximation as above, A'/T. The same trick that corrects time of fall thus gives also an improved intercept. Any harm done by using A' in place of A to determine α_1 and α_2 is slight, so A' is the only variable parameter which must be known in order to make all radar adjustments. For small vertical speed, the effect of time lags τ_s and τ_a on the altitude value to be used for the depression-angle corrections is quite negligible. Addition of τ_s to time of fall is the only important correction for radar lag.

5. LEVEL FLIGHT ROCKET FIRING

When rockets are fired forward from an aircraft in level flight, the ballistics of the motion is so complicated that an analytical approach to the problem of radar fire control seems unprofitable. Fortunately, however, a large amount of experimental data on rocket ballistics is available and an empirical solution is therefore attainable.

Fig. V.-10 shows the essential geometry of rocket firing. The rocket is initially directed along the indicated launcher line and upon ignition is subjected to a propulsive force in that direction. The wind stream tends to turn the rocket into the line of flight of the firing aircraft as soon as it leaves the launcher. The propulsive

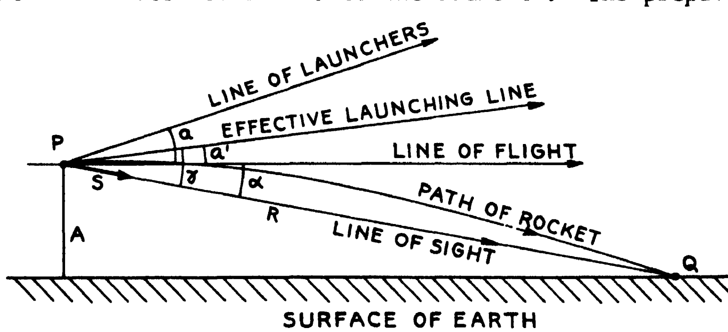


Fig. V.-10. Geometry of level-flight rocket firing.

force of the driving jet of the rocket of course changes direction as the rocket turns. The net result is that the rocket travels as though started along the *effective launching line* of the figure and not thereafter turned by the wind stream. Its propellant exhausted very soon after

launching, the rocket then falls under the influence of gravity.

Angle of attack of the rocket launchers, a in the figure, may be found from tables for any given aircraft type and launcher installation. This angle varies with gross weight and indicated air speed of the aircraft, to a degree depending upon the aerodynamic characteristics of the craft. A factor for converting from a to angle of attack a' of the effective launching line is available from other tables. This factor depends upon type of launcher and type of rocket; it varies with indicated air speed of aircraft and with temperature of rocket propellant. Still other tables provide values of the angle γ subtended by the gravity fall of the freely flying rocket. This angle depends primarily upon type of rocket and slant range to target, but is also markedly affected by true air speed of firing aircraft and temperature of rocket propellant.

From the angle of attack a' of the effective launching line and the angle γ of the gravity fall, the angular depression α of line of sight to target below the horizontal flight path follows at once by subtraction. Given slant range R to target and angular depression α, for a chosen air speed, the required altitude A for firing is determined. In low-altitude, level-flight rocket firing, all the angles involved are fortunately so small (though exaggerated in Fig. IV.-10) that no distinction need be made among angle, sine and tangent, or between cosine and unity. For low altitudes and normal rocket-firing temperatures, it is also unnecessary to distinguish between true and indicated air speed or between slant and ground range.

The available range-speed-altitude data for various rockets and aircraft can be plotted in the form of curves relating slant target range to air speed at the instant of firing, for constant altitude and propellant temperature. If wind and target motion are absent, slant closing speed is equal to air speed of aircraft. For each combination of type of aircraft, type of rocket, and type of launcher, one curve is required for each of a representative set of flight altitudes, at each of a representative set of firing temperatures. Curves that are typical of the behavior of certain fast rockets when fired from a slow aircraft

are shown in Fig. V.-11(a), while Fig. V.-11(b) presents curves typical of slow rockets when fired from a fast aircraft.

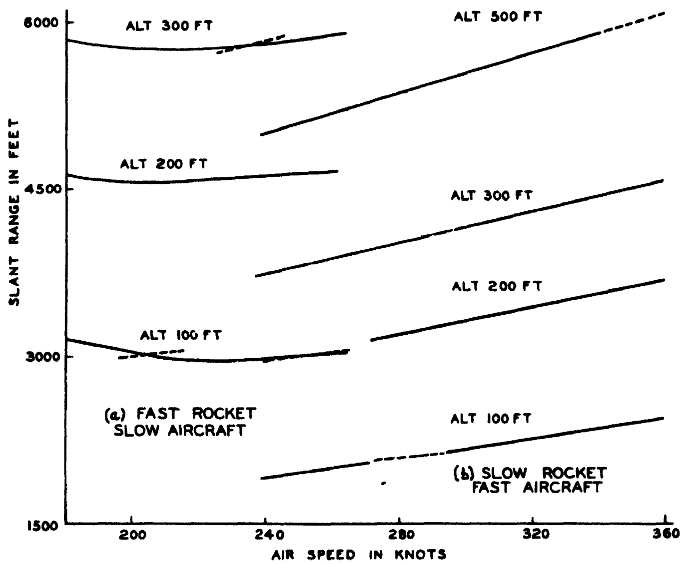


Fig. V.-11. Typical effect of air speed on rocket-firing range.

It is immediately evident from the figure that, for slow rockets and fast aircraft, a linear approximation to the range-speed relation at release will be highly satisfactory. This is the case in which the rocket turns well into the wind stream before much of its propellant is expended, so that angle of attack of the aircraft has little effect on the firing range. At the other limit, with fast rockets and slow aircraft, the direction of thrust for a considerable part of the brief interval of rocket propulsion is seriously affected by the attitude of the aircraft, and the variation of firing range with air speed is much less linear. The range of operating speeds at which any given aircraft will be used in attack is small, however. A linear approximation over a limited range of speeds in the neighborhood of rated military air speed is therefore acceptable even in the unfavorable case.

For each curve, the slope and intercept of the best approximating line may be determined graphically. From

tabulation of the constants of many such lines, it is found empirically that the approximation is hardly impaired by imposing a further restriction. This is that, for a given type of rocket, speed intercept shall always differ from the mid-range speed S_m, which is characteristic of the firing aircraft, by a constant amount S_r. S_r is therefore chosen to be independent of aircraft type, propellant temperature and flight altitude.

For each aircraft-rocket combination, it is also found empirically that the variation of approximate firing range with propellant temperature depends only slightly upon flight altitude and air speed at firing. This is of considerable practical importance, as it permits propellant-temperature corrections to be well approximated by simple constant range increments for any single combination. The temperature range to be covered is limited, as rockets burn unreliably if the propellant is too cold and may explode if it is too hot.

Because of thermal time lags, effective temperature of the propellant depends upon thermal history of the rocket for some hours before firing and must be estimated with regard to that history. Range-speed relationships for some typical effective temperature such as 60° F. may be taken as characteristic of a given rocket fired from a given aircraft, and overall behavior may then be represented by a tabulation of range increments or ballistic corrections for other temperatures.

The behavior of each combination of rocket and aircraft at the reference value of propellant temperature is fully determined to a good approximation by: the mid-speed S_m, or military operating speed, of the aircraft, the excess S_r of S_m over the common intercept of the speed-range lines on the speed axis, and the values of firing range R_m at mid speed for each value of altitude. For each combination, R_m depends only on altitude and this dependence may be plotted. Fig. V.-12 shows several curves typical of such plots, with altitude displayed on a linear scale and mid-speed range on a logarithmic scale. Altitudes used vary between a minimum safe value for blind flying of 100 feet and a maximum corresponding to a firing range of about 6000 feet.

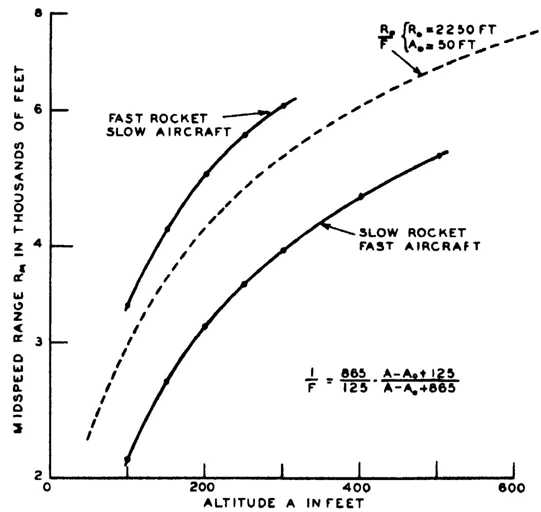


Fig. V.-12. Typical effect of altitude on rocket-firing range.

Examination of numerous graphs of R_m versus A makes it evident that all can be well approximated over an adequate range of altitude by parallel displacement of a single standard curve. This third empirical general rule is of especially great practical value. Displacement along the linear altitude scale of the figure corresponds to choice of different reference levels from which to measure altitude. Displacement along the logarithmic range scale corresponds to choice of different multiplying factors for mid-speed range. A standard curve of the general functional form

$$1/F = R_m/R_0 = \tfrac{865}{125}(A - A_0 + 125)/(A - A_0 + 865) \qquad \text{(V.46)}$$

is shown dashed in Fig. V.-12, for particular values of A_0 and R_0. A_0 is the reference from which altitude must be measured in feet to fit a specific aircraft and rocket, and R_0 is the value of mid-speed range at which to fire that rocket from that aircraft when flying at the reference altitude A_0. Both A_0 and R_0 depend only on aircraft type and rocket type. The single altitude-compensation function F given by equation (V.46) is evidently useful for both types of rocket-aircraft combination illustrated in the figure, and has in fact been found applicable as well to

all of a considerable number of actual combinations investigated.

A very large amount of ballistic data is represented to a good approximation by the group of straight lines fulfilling the equation

$$RF/R_0 - S/S_r = 1 - S_m/S_r - R_T F/R_0 \qquad (V.47)$$

between slant range R and slant closing speed S. Reference altitude A_0 and range factor R_0 depend on choice of both aircraft and rocket to be used, speed factor S_r depends only on choice of rocket, and mid-speed S_m only on choice of aircraft. Ballistic range correction R_T depends on aircraft, rocket, and departure of propellant temperature T from a chosen standard reference value T_0, while the numerical altitude-correction factor F is the function given by (V.46) of the parameter altitude-above-reference, A-A_0.

Each aircraft-rocket combination can therefore be fully characterized by a predetermined set of values for four constants A_0, R_0, S_r, and S_m, and a brief tabulation of R_T against T-T_0. Ballistic correction R_T is negative for temperatures above and positive for those below T_0, with the choice of sign used in (V.47). Firing range is fully determined by the values of these constants and the altitude A of the level firing flight.

Time lag τ_s in the radar and firing circuits was included as a range increment $S\tau_s$ in plotting the curves typified by Fig. V.-11, so is fully accounted for. Residual range R_r due to radio-frequency cables may simply be subtracted from ballistic correction R_T for propellant temperature in applying the latter, so requires no special provision. Still further range corrections may be applied in the same way as R_T, to alter the mean point of impact of the rockets so as to allow for any special operating conditions; a positive correction will move the point of impact to a position beyond the target as seen from the firing craft.

One striking characteristic of low-altitude, level-flight rocket firing may be mentioned here. This is that a slight variation in altitude of the missile upon arrival

at the target region results in a drastic variation of the range of its impact upon the land or sea surface around the target. The cause is of course the flat trajectory of the rocket and its consequent very small angle of impact. An erroneous impression of inaccuracy in level-flight rocket firing, as compared to bombing or rockets fired from a dive, can easily result from this effect.

Expressed in terms of angular depression α of line of sight, variation of which may be compared directly with that of the sighting angle used to control visual firing, the errors introduced by the approximations leading to the linear firing equation (V.47) prove in general to be quite acceptably small. These approximation errors are indeed less than the angular dispersion that is normally characteristic of rocket missiles. The presence at firing of any vertical motion of that part of the aircraft which carries the rockets will alter the aim of the line along which rocket propulsion takes place and so produce appreciable error. Rocket firing therefore requires more careful flying than does bombing. Vertical acceleration will require an altered angle of attack to maintain flight and also produce errors, though relatively small ones; the smallness of errors of this sort permits neglect of the effect of normal variations in aircraft load distribution and total loading. It may be possible to correct for measured vertical speed in an approximation of the type of (V.47), but this problem has not been investigated.

The approximation (V.47) for level-flight rocket firing differs from (V.26) for level-flight bombing in two main ways: constancy of speed intercept as altitude varies, and use of the arbitrary altitude-correction function F given by (V.46) for control of range sensitivity. Applicability of (V.47) to bombing must now be considered, to see whether one equipment can launch both types of missile. Approximation lines subject to the above two restrictions have been fitted graphically to the true bomb-release range curves of equation (V.8) (with the curves corrected for time lag τ_s) and the resulting errors have been studied. As might be expected, the result is that a satisfactory bombing approximation may still be obtained, but only over

a restricted range of closing speeds. This is not necessarily a serious limitation, as operating speed for a single type of aircraft does not vary widely, and equipment which is to fire rockets must be adjusted to the aircraft in use anyway.

The main differences found between rocket and bomb approximations of the same type are in the characteristic range R_0 and the characteristic speed S_r. Since rockets and bombs fall through equal altitude differences in roughly equal times and the rockets travel forward much the faster, R_0 is of course much larger for rockets than for bombs. For rockets, S_r is independent of aircraft speed and is of the order of magnitude of the speed increment imparted to the rocket by its own propellant. For bombs, the actual speed intercept, S_0 or S_m-S_r, is the quantity independent of aircraft speed. Different values of S_r must therefore be used when bombing from aircraft having widely different mid-operating speeds S_m. S_r need not be changed when bombing from any of various aircraft of nearly equal S_m, however. There is, of course, no propellant-temperature correction R_T for bombs. The value of reference altitude A_0 for bombs appears to be independent of aircraft. While different from that for bombs, the rocket value of A_0 varies only moderately with type of aircraft or type of rockets. A_0 does seem to depend, however, on whether rockets are fired directly from racks on the aircraft, or while in free fall after release from the racks.

6. EFFECT OF AIR RESISTANCE

a. *General.* Neglect of the effects of air resistance is quite justified in bombing from sufficiently low altitudes, at sufficiently low air speeds, with bombs of sufficiently good aerodynamic properties. What happens when these conditions are not met must now be considered. Air resistance can of course never be neglected in the case of rockets; it has been implicitly taken into account in the preceding section by use of data from actual rocket firing as the starting point.

In the absence of air resistance, it is immaterial whether speed of approach of firing aircraft to surface target is due to motion of the aircraft through the air

(air speed), to motion of the air body over the earth (wind), or to motion of the target over the earth (target speed). In the presence of air resistance, wind and target speed do not act directly to change the motion of the missile through the air, while air speed of the firing craft does so act. Effects of wind and target speed are indistinguishable, and the two together will hereafter be referred to simply as target air speed. For a given closing speed, correct missile-launching range will therefore depend on the relative values of aircraft and target air speed. F-m radar data, giving overall slant closing speed only, is thus insufficient to determine fully the correct instant for release when air resistance is important. The procedure required to allow most nearly for windage depends on the particular case in question.

b. ***Rockets.*** For rockets, the procedure of section 5 gives correctly the range in air from firing point to point of impact. The presence of a target air-speed component in the line of fire results in the target having a different position in the air at the instant of rocket impact from that which it had at the instant of rocket firing. Firing must be controlled on the basis of predicted target position at impact.

Time of flight of rockets is available from tabulated data for various rocket types, propellant temperatures, slant ranges, and aircraft air speeds. Given target air speed in line of fire, target displacement in air range during firing lag τ and subsequent rocket flight can therefore be determined. Starting at any chosen point on a range-speed curve for zero wind and target speed (see Fig. V.-11), the change in range for a small increment of target air speed may be plotted. The short dotted lines of Fig. V.-11 are such incremental target air-speed plots. Starting from the intersection of a dotted line segment and the corresponding curve in the figure, a change in closing speed due to altered air speed of firing craft will change the proper firing range according to the curve. A similar change in closing speed caused by target air speed will require the change in firing range given by the dotted line.

Inspection of Fig. V.-11(b) will show that for slow rockets fired from fast aircraft, changes in target air speed have practically the same effect as charges in air speed of firing craft, so that firing range is determined solely by overall closing speed. This happy result is probably largely fortuitous. For fast rockets fired at low air speed, Fig. V.-11(a) shows that the effect of speed variation of aircraft is strikingly different from that of speed variation of target. The best that can be done is to make a compromise choice of approximating line. Since the operating speed range of a given aircraft is much more restricted than the possible range of wind and target speeds, this choice should favor the dotted lines in slope as much as possible.

c. *Bombs.* The effect of air resistance on bombs is to cause them to lag behind or trail the dropping aircraft rather than to remain directly below it while falling, as well as to prolong slightly their time of fall. The deceleration producing the trail depends upon aerodynamic characteristics of the bomb, density of the air, and air speed of the bomb. Resulting total trail in range at impact is inversely proportional to a *ballistic coefficient* which characterizes the performance of the bomb (this is really just a definition of the ballistic coefficient), and is rather roughly proportional to the 1.7 power of the air speed. For release at zero vertical speed, trail is rather closely proportional to altitude. These well known generalizations are deduced from available tables of values of trail. Study of values of trail specially computed for dive and toss bombing has shown primary dependence on vacuum time of fall, with the separate values of altitude and vertical speed at release exerting a real but relatively slight influence on the overall result.

The best method of correcting for trail depends upon the operating conditions to be met. Where a very wide range of closing speeds must be covered without readjustment, the main variable will be air speed of bombing aircraft. For this case, the range-speed curves to be approximated may be plotted with trail as a function of speed and time of fall included, for a bomb of medium

ballistic coefficient. Trail decreases the range at which release should occur, and decreases it the more the higher the speed. This correction therefore tends to straighten out the range-speed curves of Fig. V.-4, and indeed may make the linear approximation practically a perfect one. The trail will only be correctly compensated in this way for zero air speed of target, however. Variations of radar-measured closing speed which are really caused by air speed of target will be interpreted as changes in air speed of bombing craft, and corresponding trail corrections will be provided though not needed. Yet nothing better can be done in the case of a wide speed range without introducing separate data on closing speed and air speed.

Where adjustment for a particular type of aircraft is permitted, only a very narrow range of air speed of bomb at release is to be expected. A single trail value corresponding to mid-operating speed for the aircraft concerned may be used in this case for each altitude, as a correction to release range for all closing speeds. Where it is only practicable to apply one value of correction for each altitude, the value used will normally be that for a bomb of medium characteristics. Since no wrong corrections are applied as target air speed varies, this type of operation is probably the more accurate.

As altitudes and speeds increase, the point at which neglect of trail in bombing is no longer permissible is soon reached. For all altitudes at which f-m radar has yet been found useful and all speeds at which accurate low-level bombing has been attempted, however, the trail occurring is small. This fortunate circumstance makes it unnecessary to apply extremely precise trail corrections. Cross trail, important in high-level bombing, can be neglected entirely.

7. ROCKET SIGHTING

Rockets are fired from aircraft at visible targets with the aid of an optical sight, usually from a diving approach. Firing is done when a predetermined angle exists between line of rocket launchers (see Fig. V.-10) and line of sight to the target. Sighting angle required depends upon type of launcher, type of rocket, propellant temperature, type and loading of aircraft, dive angle of aircraft in attack,

acceleration of aircraft normal to line of flight, air speed of aircraft, air speed of target, and slant range of target from aircraft. Some reduction of these ten parameters determining the sight depression is necessary to make the sighting problem at all tractable. Fortunately, type of aircraft, type of launcher, and type of rocket do not vary during the attack and can be taken into account beforehand. Normal operating variations in gross aircraft loading have a negligible effect.

If the line of aim established by the setting of the sight is continually adjusted to maintain a proper angle with the line of the launchers, the aircraft need only be flown so as to hold the target on this aiming line in order to be continuously ready for rocket firing at any preferred phase of the attack. This will result in steady flight along a smooth, slightly curved path; acceleration normal to the flight path caused by such curvature will be negligible. Additional normal acceleration may be avoided by careful flying or, if sufficiently steady, may be counteracted by a time lag in firing. Air speed of aircraft and air speed of target affect sighting somewhat differently. Their effects are similar, however, and allowance may be made for both at once, over a limited but useful range of operation, in terms of speed of closing of aircraft on target.

Sighting angle still remains a function of four variables, even under the above restrictions. These variables are range, closing speed, dive angle and propellant temperature. Depression angle of line of sight below line of launchers is, with reference to Fig. V.-10, $\alpha-\alpha'+\gamma$. This angle may be determined from tables and will be referred to as β. Dependence of β on slant range R and slant closing speed S is known empirically to be approximated closely by

$$\beta = BR - DS + JRS + \beta_1\,, \tag{V.48}$$

where B, D, J and β_1 are constants for any given dive angle, propellant temperature and type of rocket and aircraft. This approximation, however, is not well suited to the application of f-m radar.

Trial fitting of tabulated data suggests applicability

of the more convenient approximation

$$\beta = BR + D(S_0 - S) + \beta_0 , \qquad (V.49)$$

where values of S_0 and $\beta_0^{\cdot}$ are fixed for a given rocket and aircraft, but coefficients B and D still depend both upon dive angle ϕ and propellant temperature T as well. With coefficients determined by least-square fitting, this form of approximation is found to reproduce experimental data for one fast rocket fired from one slow airplane over a useful range of ϕ, T, R and S (with zero wind and target speed) quite well, though not as well as does (V.48). Root-mean-square errors in sighting angle given by (V.49) are acceptably small, and even the peak errors at the limits of R and S are less than the natural dispersion of the rockets.

Values of B and D found in this way show trends suggesting the possibility of replacing B by $(b-h\phi-jT)$ and D by $(d-pT)\cos\phi$, where b, d, j, h, and p are constants for a given rocket and aircraft. This, if acceptable, gives an approximation linear in slant range and closing speed, and with the effects of dive angle and temperature fully segregated, with seven constant coefficients for each combination of aircraft and rocket. Unfortunately, priority of other work limited this study almost entirely to a single aircraft and rocket; the general utility of the results therefore remains undetermined.

8. MOTION IN AZIMUTH

a. *Collision.* No discussion of aircraft and missile motion is complete while confined entirely to components of motion in the vertical plane through aircraft and target. Motions transverse to the line of sight are, however, of less present concern than those along that line, because f-m radar has not found any peculiarly advantageous application to measurement of transverse motion. Observation of target position and motion in azimuth by f-m radar is entirely feasible but not different in any significant way from corresponding use of other radio systems.

The important features of motion in azimuth depend upon the type of attack required. If the aircraft is to be flown to collision with its target, two main types of

approach must be considered.[1] The aircraft may be kept always pointed directly toward the target and flown in on a *homing* course. This keeps the air-speed vector of the aircraft pointed directly at the target. In the presence of wind or target motion transverse to the line of sight, the homing aircraft must turn to remain pointed at the target and so will approach along a curved path. In the other type of approach, the direct radial or *navigation* approach, the aircraft is headed steadily in such a direction as to reduce to zero the relative target motion transverse to the line of sight. This keeps the total velocity vector of the aircraft relative to the target pointed directly toward the target. The target then remains in a fixed angular position with respect to the head of the aircraft throughout the approach.

The turning radius that must be made good by the aircraft in a homing approach to a transversely moving target will depend on the closeness of approach and the ratio of aircraft speed to target speed. Detailed study of the kinematics of the motion shows that if aircraft speed exceeds twice target speed, maximum permitted turning radius must approach zero as the aircraft approaches collision with the target. Aircraft require a considerable minimum turning radius, however. This is incompatible with the vanishing maximum radius permitted in achieving collision by homing procedure at speed ratios normal for aircraft approaching surface targets.

True radial approach requires determination of the transverse component of relative motion of the target. Direct, accurate measurement of this velocity component is a problem not yet satisfactorily solved, however. A trial-and-error approximation to the radial approach is therefore used. The aircraft must be provided with a directional reference element, such as a gyroscope, which retains a fixed orientation in space. There is also provided a directional antenna or other target-sighting device rotatable about a vertical axis. In true radial approach the sighting device, when once pointed at the target and thereafter maintained pointing in the same direction in space, will continue to point at the target throughout the approach.

In actual operation, transverse motion of the target will cause it to drift out of the line of sight of the sighting device originally pointed at it. The sighting device may then be turned to point again at the target and the head of the aircraft may at the same time be turned in the same direction and by a greater amount. Both turning angles are to be measured against the fixed gyro directional reference. The sight will again drift off the target, though maintained in a fixed direction by the gyro, and the sight and aircraft may again be re-aimed. This is the navigation process required to establish approach along a definite radius from the target.

Successive corrections, with the aircraft turned more than the sight in each case, will in time reduce substantially to zero that component of the motion of the target relative to the aircraft which is transverse to the line of sight. This occurs because the heading and so the air speed of the aircraft is gradually turned away from the line of sight, developing a transverse air-speed component just sufficient to match the combined transverse components of wind and of proper target speed. Fig. V.-13 illustrates

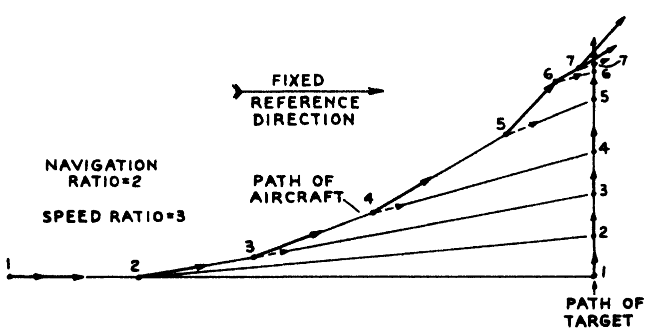


Fig. V.-13. Successive approximation to radial approach.

such an approach, as seen by a fixed observer. To simplify the picture, sight-pointing correction is made in this example only in increments of 5 degrees and only when the sight error reaches 5 degrees. After the third correction, with aircraft and target at points 4 of the figure, the approach has become radial to the full accuracy permitted by the large correction increment. The unduly frequent and large corrections just before interception result from

the excessive increment chosen for illustration. Full-line arrows in the figure represent velocity vectors of aircraft and target; dotted arrows indicate aim of sighting device.

The successive-approximation method of navigation permits satisfactorily direct approaches so long as the ratio of aircraft speed to target speed is normally high. Interception then results even when the target is maneuvering as violently as it is able to do. The ratio of angle turned through by aircraft to angle turned through by sighting device at each correction, the *navigation ratio* of the approach, has a strong effect on the path of the aircraft. When accurate sighting facilities and fast, accurate response of the aircraft to control are available, a high navigation ratio (large corrections to aircraft heading) results in rapid attainment of substantially a true radial course. For the crude sighting capabilities of a small directive antenna, high navigation ratio is only likely to introduce oscillation or hunting into the motion of the aircraft. Such undesirable effects are enhanced by the fact that the effective direction of radar-signal return from a large target at short range is often unstable, shifting at random from one portion to another of the target.

b. *Missile Launching.* Once released, a freely falling bomb is not subject to control; it remains during its fall almost directly below the bombing aircraft if the latter also is not subjected to control after release. Bombing aircraft must be well established on a radial course toward the target at the moment of release. The bombs will then continue along the radial path and so strike the target. Homing procedures do not yield bomb hits in the presence of cross winds, but navigational approach procedures do.

Rockets are aimed by turning the aircraft with its attached launching racks. A radial approach in a cross wind, with the aircraft not headed directly toward the target, cannot produce rocket hits. The aircraft must at firing be headed substantially directly toward the target, as in homing approach. The time of flight of the rockets, however, is not entirely negligible, so the aim must lead the apparent transverse target motion by a slight amount.

No work has been done toward determing this lead for rocket firing by f-m radar.

9. NOTATION AND REFERENCES

a. *Notation.* The algebraic notation used in this chapter is listed alphabetically below.

a	Vertical acceleration of aircraft (positive upward); also, angle of attack of rocket launchers.
a'	Angle of attack of effective line of rocket launching.
A	Altitude of aircraft (and radar).
A_0	Reference altitude.
A_a	Altitude at moment it is measured.
A_r	Altitude at moment bomb is released.
A_s	Altitude at moment slant range and speed are measured.
A'	Altitude modified to compensate vertical speed.
b	Partial range coefficient for rocket sighting.
B	Coefficient of range in analysis of rocket sighting.
c	Velocity of radio-wave propagation in space.
d	Partial speed coefficient in rocket sighting.
D	Difference between exact and approximate slant ranges; also, coefficient of slant speed in analysis of rocket sighting.
D_m	Maximum error in approximating slant range.
e_0	Bias component of f-m radar output voltage.
e_1	Output voltage of f-m radar at which relay is actuated.
F	Function of altitude which determines firing range.
g	Acceleration of gravity.
G	Ground range, or horizontal component of distance, between aircraft (and radar) and target.
G_1, G_2	Ground ranges at minimum and maximum horizontal approach speeds.
h	Partial coefficient of dive angle and range in rocket sighting.
h_R, h_S	Range and speed counter sensitivities in f-m radar.
H	Horizontal or ground speed of approach of aircraft (and radar) relative to target.
H_1, H_2	Limiting values of horizontal speed.
j	Partial coefficient of propellant temperature and range in rocket sighting.
J	Coefficient of range-speed product in rocket sighting.

k_R, k_S Range and speed beat-producing sensitivities of f-m radar.

L Length of radio-frequency lines in radar.

p Partial coefficient of propellant temperature and speed in rocket sighting.

R Slant range or distance between aircraft (and radar) ana target.

R_0 Reference firing range for mid speed and reference altitude.

R_1, R_2 Slant ranges at minimum and maximum horizontal approach speeds.

R_r Residual range equivalent of r-f lines.

R_T Ballistic correction to range for rocket propellant temperature.

R', R'' Range values given by linear approximations.

S Slant speed of approach of aircraft (and radar) relative to target.

S_0, S_0'' Intercept of linear range-speed approximations on speed axis.

S_1, S_2 Slant speeds at minimum and maximum horizontal approach speeds.

S_m Mean operating speed of aircraft.

S_r Reference speed, or excess of S_m over S_0.

T Time required for aircraft to travel from present position to position on vertical line through target; also, temperature of rocket propellant.

T_0 Reference temperature of rocket propellant.

T_f Time required for bomb to fall from altitude of aircraft to ground.

T' Slope of linear range-speed approximation.

v Velocity of radio-wave propagation on transmission line.

V Vertical speed of aircraft (and radar), positive upward.

V_0 Vertical speed (magnitude) gained in free fall, from altitude of aircraft to ground.

V_r Vertical speed at moment bomb is released.

V_s Vertical speed at moment slant range and speed are measured.

V_v Vertical speed at moment it is measured.

α Angular depression below horizontal of line of sight from aircraft (and radar) to target.

α_1, α_2 Angular depression at minimum and maximum horizontal approach speeds.

β Angular depression below rocket-launcher line of line of sight from aircraft to target.

β_0, β_1 Reference values of sight depression.

γ Angle subtended at aircraft by gravity fall of rocket during its flight to target.

τ_a Time lag from occurrence of measured altitude until result of measurement is available as datum.

τ_s Time lag from occurrence of final values of slant range and slant speed until bomb release.

τ_v Time lag from occurrence of measured vertical speed until result of measurement is available as datum.

ϕ Dive angle of aircraft path.

b. *Reference.*

1. L. C. L. Yuan: . "Homing and Navigational Courses of Automatic Target-Seeking Devices", Report no. 1 under contract NXsa-25337 (Dec. 13, 1943).

PART TWO

APPLICATIONS

CHAPTER VI.

SINGLE-TARGET F-M RADAR SYSTEMS

1. GENERAL

The elements described in Chapters III. and IV. may be combined and modified in various ways to produce complete frequency-modulated radar systems. For operation against single or isolated targets, these systems may indicate or control range, speed or both, or they may launch missiles under such conditions as are discussed in Chapter V.

Of the f-m radar systems which have been developed, the altimeter **AN/APN-1* (with two very similar predecessors) and the low-altitude automatic bombing equipment *AN/APG-4* reached production and saw operational use in the recent war. Production designs were completed on radar systems *AN/APG-6* (with azimuth determination) and *AN/APG-17* for automatic bombing, but these were never produced because of changing operational requirements. Rocket-firing equipment *AN/APG-17A(XN-1)* was undergoing flight tests and ready for production design when the war ended, and approach-control equipment *AN/SPN-2(XN-1)* reached a similar stage shortly thereafter. These six typical systems and their uses are described in this chapter.

Functional circuit diagrams are included to indicate circuit arrangements and component values found suitable for the systems here described, but no attempt is made to describe in detail the design and operation of all circuits. Detailed descriptions of these systems and their operation may be found in the Handbooks of Maintenance Instructions for the six equipments. Special auxiliary features which are included in the *AN/APG-6* and *AN/APG-17* equipments but are not essential for their operation are covered in Chapter VII. Methods and equipment for calibration of f-m radar systems are also discussed in Chapter VII.

System design is not a matter of routine application of a few general rules. Each system calls for the modification and elaboration of basic apparatus to meet a particular

set of specific requirements. The art of system design is therefore made up of the case histories of individual equipments. This chapter and Chapter VIII. are intended to illustrate ways in which it has seemed practicable to adapt the capabilities of f-m radar to meet highly specialized and detailed requirements. The specific military applications discussed may prove to be of only temporary interest in themselves, but have led to equipments which should remain valid examples of f-m radar system planning.

2. RADAR ALTIMETER **AN/APN-1*

a. *Description.* Altimeter models *AYB* and **AN/ARN-1* (the latter also known by U. S. Navy designation *AYL* and U. S. Army designation *RC-24*) were very similar to the **AN/APN-1* (or *AYF*). Only the last series of this line will be described; it differs from its predecessors mainly in being provided with a 400 to 4000-foot altitude range in addition to the 0 to 400-foot range common to all models.

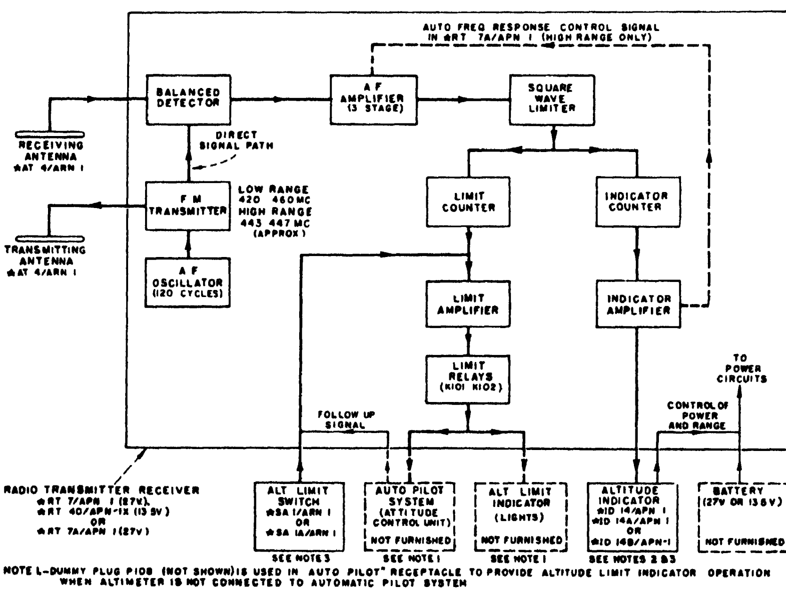


Fig. VI.-1. Block diagram of f-m radar altimeters of **AN/APN-1* series.

Fig. VI.-1 is a block diagram of the system. Starting at the middle of the left side of the diagram, sinusoidal modulating signal is developed by a vacuum-tube oscillator.

This signal drives the vibrating-capacitor modulator of a push-pull triode transmitter like that shown in Figs. III.-7 and III.-8 and described in section 3a of Chapter III. The transmitter, with a power output of one-tenth watt, is connected through a length of standard *RG-8/U* coaxial cable to the streamlined dipole antenna shown in Fig. III.-1. Another similar antenna is connected through a similar cable to a balanced detector like that shown in Figs. III.-22 and III.-23, which operates as described in section 5a of Chapter III. Local mixing signal is fed directly from the transmitter to the balanced detector over a link circuit within the equipment.

Detector beat-note output is amplified by a three-stage resistance-capacitance coupled amplifier which is unusual only in the feed-back network applied from plate to grid of its first stage. As a result of this feed back, the amplifier gain rises at 6 decibels per octave from 600 to 5000 cycles per second, as shown in Fig. III.-28, to reach a maximum at 10000 cycles per second. Gain falls off sharply at still higher frequencies, as well as at very low frequencies. In the most recent units, gain at the higher beat frequencies is automatically reduced at low altitudes to reduce interfering noise. The circuits of this entire radio-set portion of the system are omitted here, since they differ only in minor ways from those shown in Fig. VI.-7 for the corresponding parts of the *AN/APG-4*.

Fig. VI.-2 is a functional circuit diagram of the remainder of the equipment shown in the block diagram. Output from the audio amplifier drives a single pentode limiter with the bias-stable input circuit of Fig. IV.-8a. Two separate counters are fed in parallel by the limiter. One of these, a positive-output counter, is partially linearized by a cathode follower connected as in Fig. IV.-5, which serves also as a current amplifier to drive the rugged milliammeter that gives a direct quantitative indication of altitude. The discharge-path return lead of this indicator counter is tapped far down on the output resistor of the cathode follower, so that non-linearity of the relation between input frequency and output voltage is not eliminated but only reduced to the desired degree. With the meter indicator one radar system is complete.

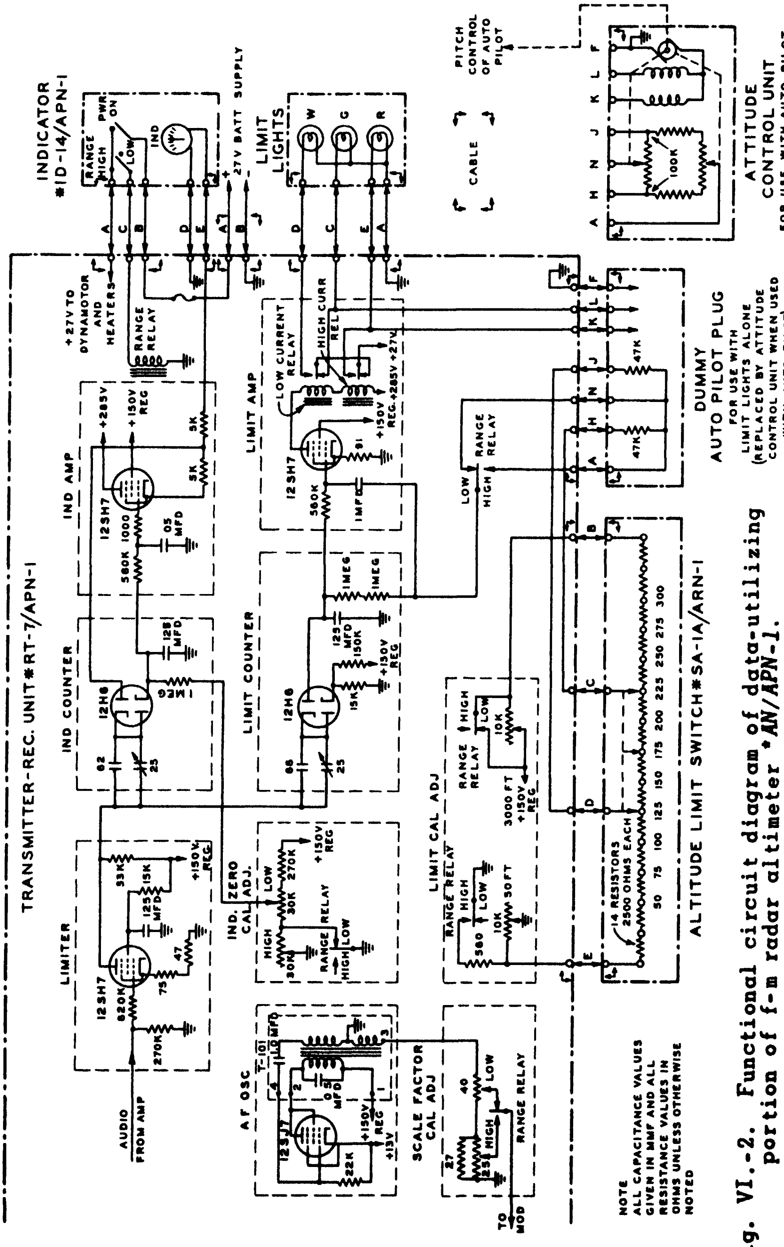


Fig. VI.-2. Functional circuit diagram of data-utilizing portion of f-m radar altimeter *AN/APN-1.

The other counter is part of an essentially separate system, which uses the same modulation, transmitter, receiver and limiter in common with the indicator system. This counter is a negative-output one of the null type shown in Fig. IV.-6. It operates relays through an amplifier, as described in section 5b of Chapter IV., when altitude departs by more than a definite limiting amount from a preset value. The preset altitude is determined by manual adjustment of an altitude-limit switch, which selects a tapping point on a voltage divider and so sets the null-output point for which the counter is balanced. The relays may actuate limit-indicator lamps or may operate a servo mechanism which flies the airplane at the selected altitude. Provision for control of flight was not made in the early *AYB* equipment. Alternatively, the manual limit switch may be omitted and the limit relays may be given full control of the counter-balancing voltage divider through a small servo, as in Fig. IV.-20; this provides an output shaft capable of adjusting other equipment in accordance with altitude. With limit switch and limit lights, or a servo-adjusted limit control, the second system of the altimeter is complete.

It may be seen from the block diagram that the complete dual system is almost entirely disposed within a single major equipment unit, designated **RT-7/APN-1*. The appearance of this transmitter-receiver unit is shown in the front and top views of Fig. VI.-3. Dust and shield-compartment covers are removed in the top view to expose the interior arrangement. At the left may be seen the radio transmitter within its shield compartment; the circular ceramic face plate of the vibrating frequency modulator is visible at the center of the floor of this compartment. The balanced detector is in the next shield compartment, while the small shock-mounted sub-chassis at the center rear carries the three-stage audio amplifier. At the right end are the dynamotor, filter capacitor and regulator tube of the high-voltage power supply, as well as the limiter tube and some calibrating rheostats. In the front center are the modulating oscillator and the counter diodes and current amplifiers of the altitude-indicator and limit-indicator circuits. The size of this 14-tube unit is indicated by the ruler included in the

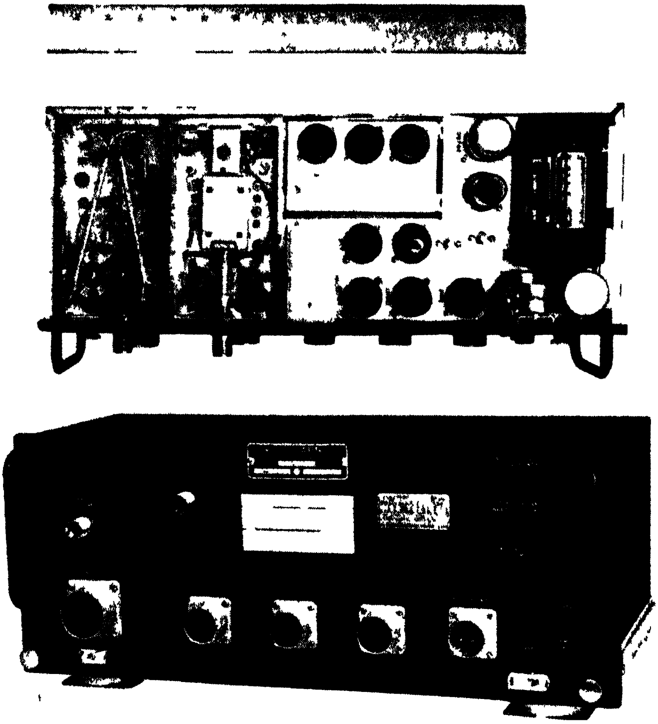


Fig. VI.-3. Main unit of *AN/APN-1* radar altimeter.

figure and its weight is 20¼ pounds, including the shock-mounting base so essential to airborne operation. Primary power required is 2.9 ± 0.3 amperes of direct current (depending upon exact condition of operation) at 27 volts (nominal), or twice that current at 13.5 volts.

Fig. VI.-4 shows the auxiliary units of the system. At the right is the milliammeter indicator, designated **ID-14/APN-1*, which may be seen to carry the on-off power switch and the altitude-range-selecting switch for the entire system. Upon throwing the selector switch to the high-range position, the scale numerals 2, 3, and 4, which are visible through windows in the dial, are replaced by 20, 30, and 40; the scale is read directly in hundreds of feet for either range. The non-linear scale marking is suited to the non-linear characteristic of the indicator counter and permits very close altitude readings to be

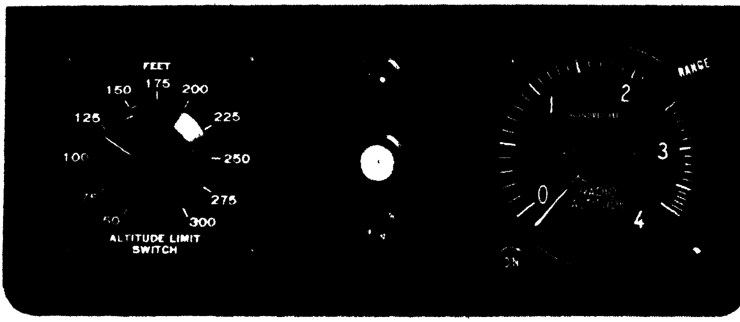


Fig. VI.-4. Auxiliary units of *AN/APN-1 altimeter system.

made when landing.

At the left of the figure is the altitude-limit switch, designated *SA-1/ARN-1, which is the altitude-setting voltage divider of the limit-indicator system. This unit permits selection of any one of the 11 integral multiples of 25 feet from 50 to 300 feet (or of 250 feet from 500 to 3000 feet on the high range) as an altitude for level flight. Limit-indicator lamps for use in maintaining level flight are shown between the other two units. The total weight of these auxiliaries is 3 pounds. Weight of interconnecting cables and plugs, always a serious item in airborne equipment, depends upon the particular conditions of each installation.

Only the power, altitude-range-selecting and flight-altitude-selecting switches shown in Fig. VI.-4 are used as operating controls. Maintenance controls for transmitter tuning, detector tuning, and detector balance are provided within the transmitter-receiver unit. Calibrating controls are provided, in duplicate for the two ranges, as twin controls under access covers on the front panel of that unit. The calibrating controls determine modulator input (hence, width of frequency band swept and altitude sensitivity of system), bias to which indicator-counter load is returned (hence indicator zero), and voltage applied to the terminals of the null-counter balancing divider of the altitude-limit system (hence, accuracy of selected level-flight altitudes). Sensitivity-calibrating controls for both counters are also provided, within the unit. Since

the main unit has no operating controls, it may be installed out of the way in any reasonably accessible portion of the aircraft that does not require excessive length of connecting cables.

b. *Uses and Operating Characteristics*. These altimeters were developed as aids to landing on water in the dark or under poor visibility, and to low-level tactical flying over water. Automatic flight control from the altimeter was developed for use in pilotless naval aircraft. Provision for servo-shaft output makes possible automatic correction of fire-control devices for altitude. An important tactical use not anticipated during development was found in maintaining accurate terrain clearance while dropping parachute troops. Another use found for radar altimeters in general is in pressure-pattern flying over water. This fuel-saving procedure involves maintaining constant actual altitude by radar, while steering through the weather pattern to maintain constant barometric pressure.

A well maintained f-m altimeter in normal operation over water manifests as its obvious operating characteristic merely the direct, continuous indication of absolute altitude. A degree of indicator damping found to give an acceptable compromise between sluggish indications and unsteady indications is used. Absolute accuracy at landing is extremely high, while errors of a few per cent may be expected at other altitudes. The obvious characteristic of limit-light operation is the altitude range over which the "on altitude" signal is given. This normally averages ± 5 feet from the preset altitude for the low range, or ± 50 feet for the high range, though use of a less sensitive relay tube will double these "dead" regions if desired.

Over very smooth water on a very calm day, indications may occasionally be seen to vary in six-foot jumps (60 feet on high range) as a result of fixed error. Over very poorly conducting land, indicated altitude will sometimes be low on the high range because of inadequate radio reflection from the ground. It is normally possible to climb to at least 8000 feet over water before the signal drops to the point of being unable to maintain a 4000-foot or full-scale indication (this loss of signal or "drop out" is largely due to

the falling gain of the audio-frequency amplifier at excessively high range-beat frequencies).

Internal characteristics not evident from overall operation are fully as important as externally evident ones. Unmodulated radio frequency is 445 megacycles per second and modulation frequency 120 cycles per second. Low-range sweep width is nominally 39 megacycles per second and is limited by the capabilities of the vibrating modulator and by undesired amplitude modulation of the radio-frequency oscillator. The resulting radar range sensitivity k_R [equation (II.22a)] is 19 cycles per second per foot. Average sensitivity of the non-linear indicator counter is $\frac{1}{150}$ volt per cycle per second. Range changing is accomplished by a relay that switches sweep width to one-tenth the above value for the high range, at the same time making slight bias changes to preserve accurate indicator-zero and altitude-limit settings; electrical damping of aircraft motion in automatic flight is also altered on changing range.

Amplitude and frequency of the modulating signal determine the scale factor of the system and must therefore be maintained accurate on an absolute basis. To accomplish this, the plate-supply voltage of the triode-connected modulating oscillator is maintained constant by a gas-discharge regulator, while modulation frequency is controlled by a factory-sealed resonant circuit. Indicator-counter sensitivity is proportional to voltage swing of the limiter, so limiter plate and screen are also fed from the regulated supply bus. The null-type limit counter is insensitive to supply-voltage variations so long as the balancing voltage divider is fed from the same supply as the limiter; the altitude-limit switch is therefore also fed from the regulated supply. Methods of calibration will be described in Chapter VII.; they are used to render correct the indications stabilized by use of regulated supply voltage.

Imperfections of operation which may sometimes be encountered are usually caused by some type of interfering noise. Microphonic disturbances, particularly in the balanced detector and audio amplifier, were only with difficulty reduced to a harmless level; this is the normal state

of affairs in high-gain electronic equipment used under the conditions of extreme vibration customary in military aircraft. Good maintenance to prevent occurrence of faults in tubes, connections, or shock mountings is necessary to insure that the microphonic level will remain low. Radio-frequency cables and connectors are particularly prone to produce microphonic noise, as are loose elements of the external aircraft structure in the field of the antennas. Especially for high-range operation, great care in locating and making the antenna installations is necessary to avoid field modulation or other noise as well as to minimize direct feed through of signal.

High-frequency noise, emphasized by the response characteristic of the audio amplifier, is most likely to be troublesome at the low-altitude end of the high range, where the desired-signal frequency is rather low and signal strength is also moderately low. The final modification of the **AN/APN-1* series incorporates altitude-operated automatic gain control, of a type which reduces high-frequency response at low altitudes without markedly affecting low-frequency response. Another type of noise which must be kept low by good maintenance results from detector unbalance. Detector balance is perfect at not more than two or three frequencies in the band swept during modulation, and balance difference across this band gives rise to spurious output at the modulation frequency and its harmonics. Normally good balance results in reduction of sensitivity to amplitude modulation, even at the worst-balanced frequency in the band swept, by a factor of at least five relative to a single detector.

Extremely low beat frequencies are never encountered in operation, even when landing, because of the residual altitude contributed by the radio-frequency transmission lines between transmitter-receiver unit and antennas. The equipment is adaptable to widely varying antenna installations, with total residual altitudes varying from 13 to 58 feet, and must be calibrated to match the installation with which it is to be used. Coupling to the transmitting antenna must be quite loose if disturbance of modulation-sweep calibration by antenna mismatch is to be avoided.

Beside the altitude equivalent which is resident in the

r-f lines, there is that due to the remaining air path from transmitting antenna to ground to receiving antenna when the aircraft is in flying position with its wheels touching the ground. Because of the lateral separation between antennas and resultant obliquity of air paths to and from the ground-reflection point, this air-path residual altitude exceeds the actual average altitude of the antennas at landing. On the other hand, at high altitude the transmission paths are practically vertical and the air-path residual is the actual average of the height of the two antennas above the wheels. Variation of residual altitude with true altitude has been called "mushing error"; its effect may be minimized by a trick of calibration.

A number of military uses for radar altimeters became evident at about the time that technical development had reduced bulk and weight of f-m equipment to a point permitting its use even in fighter aircraft. Introduction of the balanced detector and of internal mixing-signal coupling from transmitter to receiver gave at the same time a great improvement in performance. This improved performance, together with reduced bulk and weight and the emergence of military needs, accounts for the very wide use of the **AN/APN-1* and its immediate predecessors as contrasted to purely experimental use of other radar altimeters developed earlier. The f-m altimeter is one of the few war-tested radar devices that is likely to find immediate and continuing use in peace-time aviation.

c. ***Automatic Flight Control.*** Automatic flight under Altimeter Control of Elevators (*ACE*) using the **AN/APN-1* is accomplished by three servo mechanisms operating in cascade. This is such a complex dynamical system that no attempt at a complete analysis will be made here. It is also an excellent example of the ease with which f-m radar can be integrated with other control equipment, so merits full description. Some geometric terms to be used in this description will first be defined and their relations to other factors indicated.

Angular position or orientation of an aircraft is called its *attitude* and can be fully specified by three angles. Rotation of the aircraft about a transverse axis through its center of gravity (parallel to the wing span) is called

pitch; measured from the angular position the aircraft maintains when in level flight, this rotation is the *pitch attitude* of the craft. This is the angle of immediate interest. Neglecting the small operating variations of angle of attack of the aircraft to the air stream, the flight path of the aircraft tilts just as the craft pitches. Rotation about the longitudinal axis of flight is called *roll*, or *bank* when measured from the normal wings-level attitude, and is not of present concern. Rotation about a vertical axis of the vertical plane in which the longitudinal axis of the aircraft lies is called *yaw;* measured from the vertical geographic-meridian plane, this rotation angle is the *heading* of the aircraft. Heading and yaw will be of importance in the later discussion of the operation of the *AN/APG-6* equipment.

Rotation of the aircraft in pitch is controlled aerodynamically by moving its *elevator* airfoils and rotation in yaw primarily by moving the *rudder* airfoils; roll is controlled primarily by *ailerons*. Effects of elevators and rudder are interchanged for large angles of bank, but that is of no concern in normal automatic flight. Angular setting of each control airfoil is a measure of time rate of rotation of the aircraft about the corresponding axis. Pitch attitude, corresponding to the direction of flight in the vertical plane, is a measure of rate of change with time of aircraft altitude. These rate relations are important in the behavior of servo mechanisms, because introduction of a control action proportional to rate of change can produce the effect of viscous damping of the mechanism without increasing inertia. Higher-order derivative controls are also used sometimes to improve operation of servo mechanisms, but need not be considered here.

Fig. VI.-5 is a block diagram of the complete *ACE* system for automatic flight at constant terrain clearance. Elements which produce as output the algebraic sum of two input data are indicated as differentials. Elements which produce a sensed control-actuating output by comparing two input data and determining their difference are indicated as control comparators. Solid lines represent electrical and dashed lines mechanical interconnections. Departure of the aircraft from a preset altitude is made,

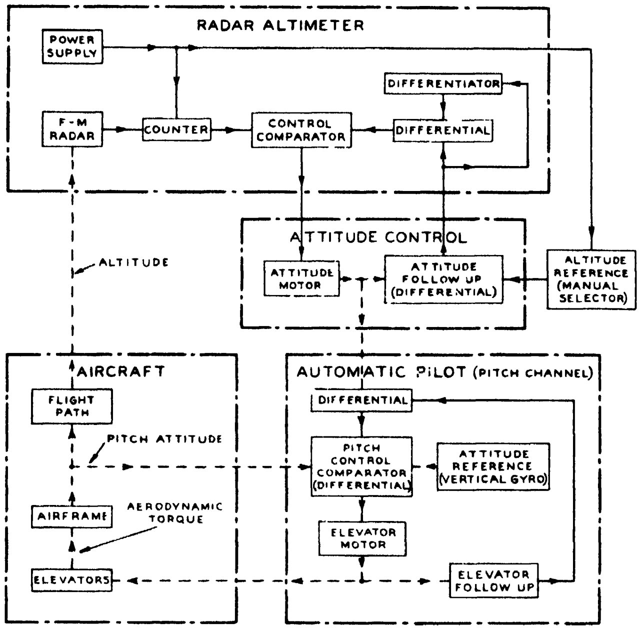


Fig. VI.-5. Block diagram of automatic control of flight altitude by radar altimeter.

by action of the altimeter and an associated attitude-control servo, to adjust proportionally the pitch attitude in which the automatic pilot is ordered to fly the aircraft. Departure of the aircraft from the pitch attitude in which the automatic pilot is set to fly it is made, by action of the pitch-control channel of the pilot, to adjust proportionally the position of the elevator surfaces. Departure of the elevators from their neutral position is made by the aerodynamic action of the elevators to start the airframe rotating in pitch, while departure of the airframe from the zero-pitch attitude of horizontal flight is made by the aerodynamic action of the wings to start the aircraft moving vertically. It should be noted that the aerodynamic actions of the third servo mechanism (the aircraft itself) establish rates of motion only, while the control actions of the other two servos establish total motions directly.

The block diagram shows five different feed-back paths, including the overall composite-servo loop. Two of these

paths serve a dual purpose. Additional feed-back paths representing aerodynamic stability and damping of the aircraft, as well as possible damping of elevator-control motion, are omitted. This multiplicity of feed-back loops indicates the inherent complexity of the complete system. Only the fact that some elements of the system are fundamentally much slower in their action than others makes overall behavior amenable to simple description.

Action of the altimeter and its associated attitude-control unit is inherently rapid. Action of modern automatic pilots in adjusting position of elevators is also inherently rapid. Operation of these subsidiary servo loops is therefore relatively little affected by their incorporation in the larger system and may be considered simply and separately. The rate-controlled motions of the aircraft in pitch and ascent or descent are inherently slower and may be considered as under instantaneous control by the other elements.

Altimeter and attitude-control unit act in the following way. The limit-counter circuit of the altimeter impresses the voltage drop in the counter load due to counter-output current, and an altitude-reference voltage set by the manual altitude-limit switch, differentially on a control comparator consisting of the limit-relay amplifier and relays of the altimeter. The limit relays, through other relays in the attitude-control unit, operate a motor which turns an attitude-control shaft bearing a follow-up potentiometer. The follow-up potentiometer is connected as a differential on the altitude-limit switch, so that voltage variations due to shaft motion add to the manually set reference voltage fed into the comparator. The motor shaft therefore seeks and holds an equilibrium position proportional to the departure of the aircraft from the preset reference altitude. A resistance-capacitance circuit fed by the follow-up voltage acts both as a differentiator to produce a voltage proportional to rate of shaft rotation and as an electrical differential to add this voltage to the reference and follow-up voltages applied to the comparator. Rate-voltage feed back to the comparator provides viscous-type damping of the servo action. Aircraft attitude does not affect this part of the system at

all, and altitude can not change appreciably while the attitude-control servo is reaching equilibrium.

There have been and are many varieties of automatic pilot, and no single description can cover accurately the operation of all of them. Whether powered pneumatically, hydraulically or electrically, and interconnected for control by cables, hydraulics or electrical circuits, they are basically similar in principles of operation, however. Automatic pilots have three control channels, corresponding to the three aircraft attitude angles; only the pitch-control channel is of present interest. Actual pitch attitude of the aircraft is compared by the auto pilot with an attitude reference provided by a vertical-axis gyroscope maintained erect by average gravity, and departure from reference pitch causes a motor to move the elevator surfaces of the aircraft.

Elevator motion is monitored by a follow-up element and added in suitable proportion by a differential to actual pitch attitude before comparison with reference attitude. Elevator position therefore seeks an equilibrium proportional to pitch error. Effective reference pitch can be altered by adding an external control angle, in the present case the position of the attitude-control shaft, through another differential (usually in the actual-pitch channel). If necessary, elevator motion may be damped by derivative feed back (not shown). In modern auto pilots, mechanical follow-up data is likely to be fed back by electrical synchro, external control to be fed in by differential synchro, and actual pitch to be both fed in and added to follow up and control by a synchro transformer with stator attached to the airframe and rotor to the gyro. This transformer will also be the input element of an electronic control comparator actuating an electric elevator-driving motor.

The auto-pilot action is rapid, so may be considered as adjusting the elevators continuously in proportion to the actual airframe-attitude error, with respect to a reference established jointly by gyro vertical and the pitch-control signal fed in from the altimeter servo. The latter signal is proportional in turn to altitude error of the aircraft. This rapid action may be regarded as independent of

aircraft motion. As the rate of change of pitch induced by the elevators alters the actual pitch attitude of the airframe toward the altitude-error-controlled reference attitude, the auto pilot will reduce proportionately the airframe-controlling displacement of the elevators. The elevator follow up thus serves a second purpose, providing rate-of-change-of-pitch feed back to the comparator and thereby effecting viscous damping of the motion of the aircraft in pitch. The aircraft is made to fly smoothly in the pitch attitude dictated by the altimeter, which is proportional to altitude error.

As the vertical motion caused by controlled departure of the airframe from level-flight attitude changes actual aircraft altitude toward the chosen reference altitude, the altimeter attitude-control servo reduces its demand for pitch departure, by virtue of its attitude follow up, and the auto pilot in turn demands reversed elevator action. The aircraft therefore approaches horizontal-flight attitude as it nears reference altitude. The attitude-control follow up of the altimeter servo thus also serves a second purpose, providing rate-of-change-of-altitude feed back and thereby viscous damping of the motion of the aircraft in altitude.

The aircraft can not follow in altitude, and probably not even in attitude, all details of the fast operation of the altimeter attitude-control servo. However, the dead space of the altimeter limit-circuit relays often results in altitude changes near reference altitude being made as a series of small steps. If the altimeter servo is allowed to oscillate by omitting or reducing its own derivative-damping circuit, or better by externally forcing a fast small-amplitude oscillation, aircraft motion will not be directly affected but the servo dead space will be elimininated and very smooth flight control assured. To avoid momentarily dangerous flight conditions upon calling for a sudden large change in reference altitude, the range of attitudes that the altimeter servo can demand must be definitely limited to a safe value.

Some older auto pilots utilize the elevator surfaces of the aircraft not only to control rate of change of pitch attitude but also to supply steady aerodynamic lift to

compensate changes of flight trim resulting from variation of total load or load distribution in the aircraft. Elevator position, aircraft attitude and attitude-control input are added together in control of the auto pilot; flight in the normal attitude necessary for a horizontal path, but with elevators offset to maintain trim, therefore requires that the attitude control be correspondingly offset from its neutral position. To accomplish this control offset by way of the altimeter, the aircraft actually flies at an altitude slightly different from the reference value for which the system is set. This error is no fault of the altimeter control but represents a shortcoming of particular auto pilots; when it is present and objectionable, special auxiliary means must be used to remove it.

If conditions of relative response rate in various portions of the system are not the simple ones here assumed, complicated interactions can take place. These may produce peculiar and hard-to-remedy types of overall instability, but may be avoided by taking care that the assumed conditions do exist. Action of auto pilot and aircraft is rather involved in detail, but sufficient development has been done by designers of auto pilots and aircraft so that over-all operation is now very satisfactory; the internal difficulties need not bother the user.

It should be noted finally that adaptation of the f-m radar altimeter to control of level-flight altitude is in itself very simple, given a smoothly operating system of automatic pilot and aircraft. Addition of the limit circuit to a meter-indicating altimeter required only two tubes, two relays, a reference-altitude switch and a few resistors and capacitors. Provision of attitude-control input for an auto pilot from the limit circuit further requires only a small motor with reduction gearing (and optionally two more relays), and a potentiometer, as shown in Fig. VI.-2. Of course, the characteristics of the attitude-control servo must be made compatible with those of the particular auto pilot with which it is to be used. Operation of a well integrated *ACE* system is very smooth, rapid and accurate. It will, for example, cause the aircraft to follow terrain-level fluctuations accurately, within the attitude limitations imposed for safety.

3. LOW ALTITUDE AUTOMATIC BOMBING EQUIPMENT *AN/APG-4*

a. *Purpose and Description.* Light-weight, fully automatic equipment was required for accurate bombing from low-flying pilotless aircraft, or from small manned aircraft attacking surfaced submarines or other vessels at night and in low visibility. In view of the good results obtained in automatic flight control with the f-m radar altimeter, development of an f-m radar system for automatic low-altitude bombing, designated *AN/APG-4* and often referred to as the *Sniffer*, was undertaken to meet the above requirement. The usual rule that antenna design has a strong influence on radar system development applies to the Sniffer, which was first developed at a frequency permitting the direct use of the Yagi antenna array designed for the *ASB* series of pulse search radars. A scaled version of the same antenna, shown in Fig. III.-3, was used in the final 410-megacycle production equipment.

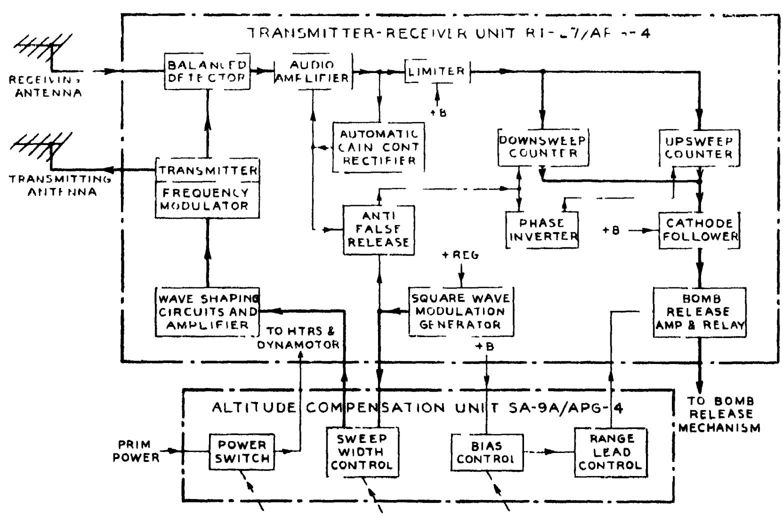


Fig. VI.-6. Block diagram of radar equipment *AN/APG-4* for automatic low-altitude bombing.

The equipment is organized in accordance with the block diagram of Fig. VI.-6. Modulating voltage from a square-wave source is adjusted in accordance with flight altitude to give the proper width of frequency sweep, and is then

fed to a wave-shaping circuit of the sort described in section 4c of Chapter III. The shaped voltage excites a modulating amplifier which drives the vibrating-capacitor modulator, which in turn swings the frequency of the one-fifth-watt push-pull transmitting oscillator described in section 3a of Chapter III. The transmitter feeds the Yagi antenna array of Fig. III.-3 through a flexible coaxial transmission line, and in addition supplies mixing signal directly to the balanced-detector receiver described in section 5a of Chapter III. Received signal is fed to the balanced detector through a similar line from another Yagi array. Beat-note output from the detector is amplified by a three-stage, high-gain audio amplifier, made selective by feed back in the first stage and by shunt capacitors in later stages. Automatic gain control of the feed-back stage, working from the rectified amplifier output, operates with strong signals to reduce gain above the peak-response frequency and to move the peak to lower frequency, as in Fig. III.-28, without markedly affecting low-frequency gain.

A complete circuit diagram of the *AN/APG-4* equipment, arranged according to function, is given in Fig. VI.-7. Differences between *AN/APG-4* and **AN/APN-1* in the portions of the circuit described above are not great. In the altimeter, the output transformer of the modulating stage is resonated to the desired modulation frequency and provided with a feed-back winding, so that the modulating tube of Fig. VI.-2 functions as a self oscillator and does not require the external drive shown in Fig. VI.-7. A high resistance in the cathode circuit of the transmitter oscillator, in the Sniffer only, develops sufficient bias to prevent oscillation and permit observance of radio silence. When transmission is required, this resistor is shorted by an external switch and oscillation occurs. The mixing-signal coupling loop in the detector of the altimeter does not have the highly symmetrical form shown in the Sniffer circuit. The audio amplifier of earlier altimeters did not have the high-ratio input transformer, grid and plate shunt capacitors, automatic gain control, or overall voltage gain of 500,000 required in the *AN/APG-4*, and its first-stage circuit constants were chosen to produce a somewhat broader, flatter gain-frequency characteristic. In the most recent altimeters, transformer input to the

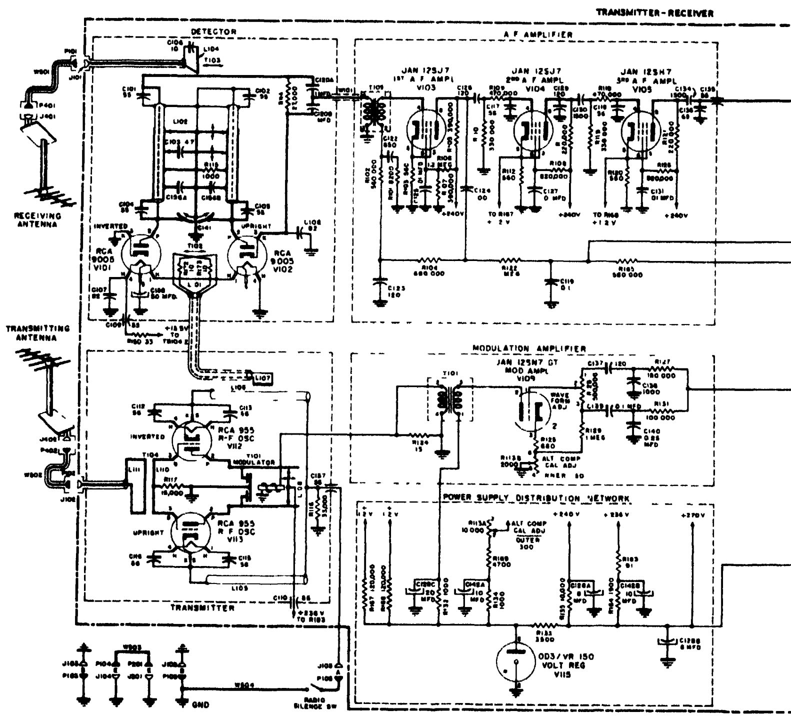
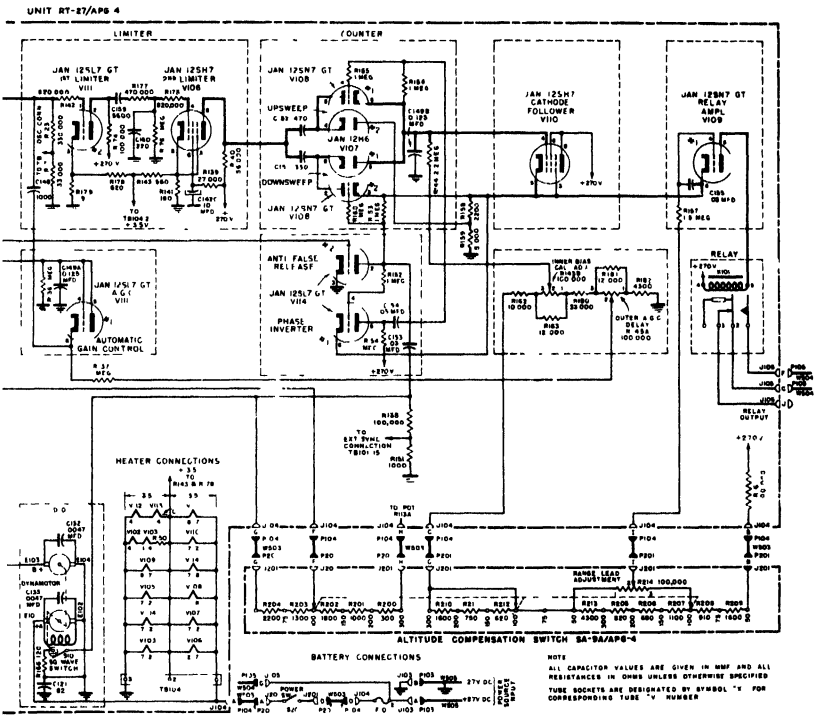


Fig. VI.-7. Functional circuit diagram of f-m radar

audio amplifier is used and frequency-dependent automatic gain control is applied on the high-altitude range only; gain control is however based on altitude rather than on signal level and is obtained from the cathode voltage of the altitude-indicator amplifier.

A noise-reducing double limiter of the sort described in section 2f of Chapter IV. is driven by the output of the audio amplifier of the *AN/APG-4*. Square-wave signal output from the second limiter actuates both a positive-output counter, active only during the frequency upsweep of the modulation cycle, and a negative-output counter of less sensitivity, active only during the downsweep. These counters have a common load and are linearized by a common cathode follower, as shown in Fig. IV.-12 and described in section 3b of Chapter IV. They provide a net d-c output having a positive component proportional to range and a negative component proportional to speed. This output, at



equipment *AN/APG-4* for automatic low-altitude bombing.

the low impedance of the follower cathode, actuates a relay amplifier connected as in Fig. IV.-19 and supplied with grid voltage which is adjusted in accordance with flight altitude. Operation of the relay causes release of a bomb at the correct range for the speed and altitude used. Bomb release is accomplished through the normal equipment of the aircraft.

Instead of using the transformer of Fig. IV.-12, the downsweep counter is synchronously switched into and out of operation by direct capacitive coupling to the square-wave modulation source, while the upsweep counter is switched in opposite phase by a phase-inverting amplifier. In the absence of adequate signal, an anti-false-release tube short circuits the counter-switching voltage, and so disables the negative counter and positively prevents the cathode-follower output from falling to the release point. Negative a-g-c voltage in the presence of a good

signal cuts off the anti-false-release tube and permits normal operation of the switched counters.

Adjustment for attack altitude is made in a small, separate altitude-compensation unit arranged to permit manual selection of any chosen one of six fixed altitudes between 50 and 300 feet. This selects the proper tap on the voltage divider controlling the amplitude of the square-wave modulating signal (r_1, r_2 of Fig. III.-12), as well as selecting the bias applied to the relay amplifier. Bias setting is controlled by two switch-type rheostats ganged with the modulation divider, one to compensate for speed intercept of the bombing approximation used and for residual range of a standard r-f cable installation (see section 3b of Chapter V.), and the other to control the voltage across a "range lead" potentiometer. This potentiometer, also located in the altitude-compensation unit, is used to set the distance by which it is desired that the bomb should fall short of the target.

The one-pound altitude-compensation switch *SA-9A/APG-4*, built in a standard A-N aircraft instrument housing, is shown in Fig. VI.-8. Connection is made by cable and plug to a single receptacle at the rear of the unit. The on-off

Fig. VI.-8. Altitude-compensation switch of low-altitude bombing system.

power switch for the radar system is included in this unit and actuated by the same knob as the altitude-selector switch.

Except for the two antennas and the altitude-compensation switch, the complete *AN/APG-4* system is contained in a single main unit (*RT-27/APG-4*), very similar mechanically to and of the same size as the main unit of the **AN/APN-1* altimeter. Fig. VI.-9 shows a complete low-altitude radar bombing system, for operation at manually selected fixed altitudes.

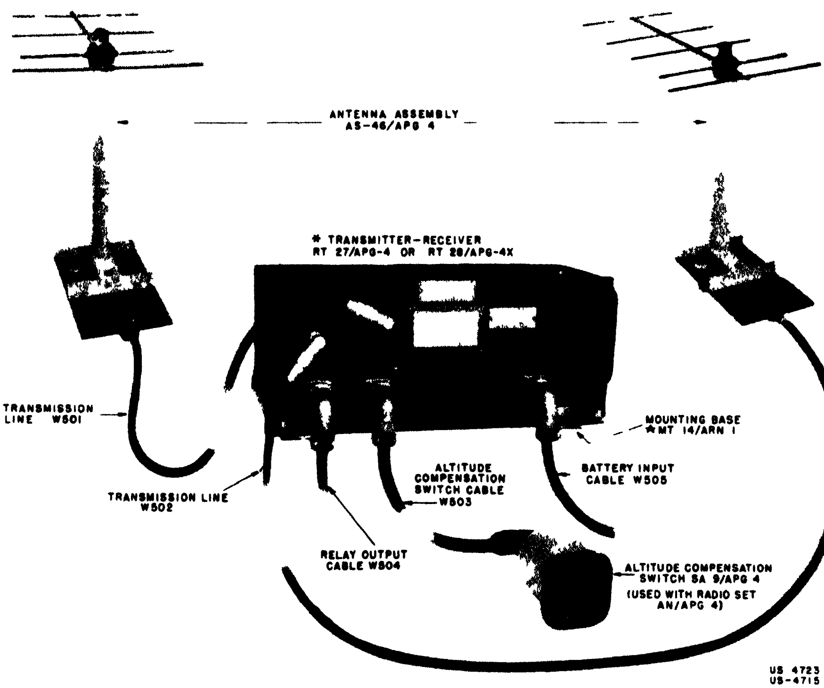


Fig. VI.-9. *AN/APG-4* radar system for automatic low-altitude bombing.

Ranging accuracy depends upon amplitude of modulating signal; the square-wave modulation generator and the modulation amplifier are therefore supplied with power at a regulated voltage. The counter system is self-compensating for variation of supply voltage; limiter plate, counter-load return voltage, and relay-tube bias can therefore be supplied from a common source of unregulated voltage. The square wave for driving the modulator and switching the counters is derived from the regulated direct voltage supply by a fast-acting mechanical switch, which is mounted on the dynamotor frame and actuated by an eccentric cam on the dynamotor shaft. Very careful damping of this switch

is necessary, and by proper choice of cam material, cam wear is made to compensate contact wear and maintain accurate switch timing over a considerable operating life.

The power switch, altitude selector and range-lead control are the only operating controls for the system; they are all located on the altitude-compensation unit. There are no operating controls on the main transmitter-receiver unit, but within this unit are maintenance controls for adjusting modulation-sweep linearity, transmitter tuning, detector tuning and detector balance. Accessible through holes in the front panel are screw-driver adjustments for calibrating sweep width, altitude compensation, and release bias, as well as for setting the signal threshold for a-g-c and anti-false-release operation. Front-panel receptacles are provided for connection to the transmitter-receiver unit of cables from the primary power source, transmitting and receiving antennas, altitude-compensation switch, and bomb-control circuits of the aircraft.

b. Operating Characteristics. Since the *AN/APG-4* was developed for use in pilotless aircraft or in small craft where operation by the pilot himself may be necessary, the operating procedure has to be very simple. Upon establishing contact with a target, or if possible a few minutes earlier, the equipment is turned on and allowed to warm up while maintaining radio silence. The altitude-compensation switch is set for that level between 50 and 300 feet at which it is desired to attack, and approach to the target is begun. When within about three-fourths mile of the target, the aircraft is leveled off at the preset altitude with the aid of the radar altimeter and is so headed, on the basis of any available aiming data, that it will pass directly over the target. No aiming information is provided by the *AN/APG-4*. Transmission is then initiated by closing a radio-silence switch, and the bomb-release system of the aircraft is armed.

At the proper point, the bomb is released automatically; the system is then inactivated manually and withdrawal or preparation for a new approach is begun. The pilot in order to attack has only to choose his altitude, activate the system and fly straight and level at the chosen altitude

and in the correct direction; the rest is automatic. Of course, if a bomb impact short of the target is desired, for example to straddle the target with a train of bombs, the proper "range lead" may be set in on the compensation-switch unit.

With a well maintained system, the characteristics observed in normal operation are rather those of the target and its surroundings than those of the radar system, and are correspondingly complex. The great value of oscilloscopic observation of beat-note signal during tests that involve actual airborne approach to a target, as a means of assessing operating characteristics, should be noted. Sufficient test drops of bombs have been made to show clearly that altitude and speed compensation both occur as intended and that no significant systematic errors are present in a carefully calibrated system. In the case of a large, complex target such as a ship, it is evident that the "radar center of gravity" of the target will shift about as the relative contributions of the highly directive reflections from various parts of the target structure vary in magnitude and phase. Photographically simulated bomb drops on ships confirm this instability of radar aiming point under broad-beam irradiation from the *AN/APG-4*. Although the equipment was designed for operation only at altitudes up to 300 feet, very limited tests[1] have shown fair operation up to 800 feet.

Fading signals, caused by high directivity of extended or complex targets, can interfere with accuracy of release. If the noise during a fade is of high frequency, as in the case of distant sea-return signal or of modulation, due to loose parts of the aircraft, of the local field which couples the two antennas, release may be delayed. If the noise during the fade is of low frequency, as in the case of altitude signal or detector unbalance, premature release may occur. Use of the integrating double limiter does much to prevent delayed release, though increasing the chance of premature release. The anti-false-release circuit usually prevents premature release. The anti-false-release threshold is adjusted in flight, at normal attack altitude but without a target, until false release is just avoided in the most violent maneuvers used in the latter part of an approach. The threshold setting,

and consequently the size of the smallest reliable target, will depend upon the condition of the sea surface when the threshold is set.

An absolutely essential requirement for acceptable operation is that noise throughout the system must be held to a practical minimum. Rough sea makes reliable operation on very small targets impossible, since the equipment is not capable of determining whether signal for a given range is coming from the target or from the sea surrounding the target. Noise extraneous to the system may come from such sources as sea return, field modulation, mechanical vibration, and electrical interference from other equipment. In the absence of all extraneous noise, properly operating equipment will tolerate a power loss of 75 decibels from the transmitter-output receptacle to the receiver-input receptacle of the *RT-27/APG-4* unit, and still provide accurate automatic release. This includes losses in transmitting and receiving r-f antenna lines, as well as the losses in radio transmission and target reflection discussed in section 5 of Chapter II. An absolute minimum echoing area of usable target is determined by this maximum tolerable loss; practically, however, some extraneous noise is always present to lower the tolerable signal loss and increase the size of the minimum accurately usable target. Reduction of microphonic and other excess noise is of utmost importance in the use of this equipment.

The primary internal characteristic of the equipment is its ability, developed as described in Chapters II. and IV., to actuate a relay when range and speed of the aircraft relative to an isolated surface target are so related that

$$k_R h_R R - k_S h_S S = e_1 - e_o . \qquad \text{(VI.1)}$$

k_R and k_S are sensitivities of the radar in converting range and speed respectively to beat-note component frequencies, while h_R and h_S are counter sensitivities converting range and speed frequencies respectively to component direct output voltages. e_o is voltage at the cathode-follower grid in the absence of beat-note signal, determined by bias voltage applied to the counter

load, while e_1 is follower-grid voltage at which the relay operates, determined by bias applied to the relay tube. k_R is controlled by product of sweep width and modulation frequency, and k_S is controlled by radio carrier frequency. h_R and h_S are determined by effective counter-input voltage swing E'_e, counter-capacitance values, and counter-load resistance. Relation (VI.1) between range and speed for release-relay operation is graphically a straight line, with slope $k_S h_S / k_R h_R$ and intercept $(e_o - e_1)/k_S h_S$ on the speed axis.

Bombs will strike approximately at the target, as shown in section 3 of Chapter V., if released in level flight when range and speed are so related that

$$R/T' - S = (R_r + R_1)/T' - S_o. \tag{VI.2}$$

Graphically, this is also a straight-line relation, with slope T' and intercept $S_o - (R_r + R_1)/T'$ on the speed axis. T' and S_o are completely determined by flight altitude, operating-speed range of the equipment, and time delay in operation, according to equations (V.23), (V.24), and (V.28). R_r is a fixed residual range corresponding to time delay in propagation through the radio-frequency lines to the antennas, and R_1 an adjustable distance or "range lead" by which bombs may be made to fall short of the target if desired.

Comparison of (VI.1) and (VI.2) shows that the *AN/APG-4* will bomb accurately if its sensitivity ratio is made to equal the slope time of the bombing approximation, and its output-voltage change due to radar signal is made to correspond to the bombing-approximation speed intercept, for the actual altitude of operation. The required correspondence is achieved in this equipment by adjusting according to altitude the frequency-sweep width in modulation and the bias applied to the relay tube. Radar speed sensitivity k_S is fixed by the radio-frequency channel used, and values for k_R, h_R, and h_S are chosen to give convenient beat-note circuit constants. Numerical values encountered in actual design and operation of the *AN/APG-4* may serve to lend concreteness to this discussion and that of earlier chapters; such values are given in Tables VI.-1 and VI.-2.

TABLE VI.-1

Design Characteristics of *AN/APG-4* Equipment

General	
Power Supply:	27 volts (nominal) d-c
Power Consumption:	65 watts (2.5 amperes)
Number of Tubes:	15
Weight complete with Antennas and Shock Mount (but less Cables):	37 pounds
Radar Portion of RT-27/APG-4 Unit	
Radio Carrier Frequency:	410 megacycles per second
Radar Speed Sensitivity k_S:	0.834 cycles per second per foot per second (1.41 cycles per second per knot)
Modulation Frequency:	110 cycles per second
Usable Frequency Sweep:	1 to 5 megacycles per second
Usable Radar Range Sensitivity k_R:	0.45 to 2.25 cycles per second per foot
Permissible Attenuation in Transmission Path from Transmitter to Receiver (no external noise):	At least 70 decibels
Audio Amplifier Voltage Gain (peak):	500,000
Audio Frequency Response:	Rises 8 decibels per octave from 600 to 5000 cycles per second, falls rapidly below 400 and above 5000 cycles
Altitude-Compensation Switch SA-9A/APG-4	
Flight Altitude:	50 to 300 feet, six fixed values
Closing Speed:	100 to 350 knots (approx. 170 to 600 feet per sec.)
Time-Lag Allowance:	0.40 second total (0.32 second in circuits, 0.027 second in relays, 0.053 second in release mechanism)
Residual-Range Allowance:	50 feet (equivalent of average r-f line length)
Range Lead:	Zero to 100 feet
Computing Portion of RT-27/APG-4 Unit	
Audio Input for Limiting:	Greater than 7/8 volts r-m-s.
Effective Counter-Input Voltage Swing:	255 volts
Modulator Lag:	4½ per cent of full modulation cycle
Counter Range Sensitivity h_R:	0.0405 volts per cycle per second

Counter Speed Sensitivity h_S:	0.190 volts per cycle per second
Overall Speed Sensitivity $k_S h_S$:	0.158 volts per foot per second
Bias Applied to Counter Load:	58 volts (approx.)
Increment Caused by Extra Counts:	9 volts
No-Signal Voltage at Follower Grid.	67 volts (approx.)
No-Signal Voltage at Follower Cathode:	71 volts (approx.)
Relay-Amplifier Grid Voltage for No-Signal Release:	63 volts (approx.)

TABLE VI.-2

Typical Operating Conditions

Approach Kinematics	
Altitude:	300 feet
Time of Fall of Bomb:	4.32 seconds
Approximation Slope, T':	4.55 seconds
Slant Speed (mid-range):	320 feet per second (187 knots)
Approximation Intercept, S_o:	-26.7 feet per second
Residual-Range Speed Intercept:	-11.0 feet per second
Range Error of Approximation:	+10.7 feet
Slant Range at Release:	1628 feet
Range Lead:	Zero
Conditions in Equipment	
Relay-Amplifier Grid Voltage:	69 volts (approx.)
Width of Frequency Sweep:	1.92 megacycles per second
Radar Range Sensitivity k_R:	0.860 cycles per second per foot
Overall Range Sensitivity $k_R h_R$:	0.0348 volts per foot
Speed Frequency:	267 cycles per second
Range Frequency at Release:	1400 cycles per second
Range-Counter Output:	+56.7 volts
Speed-Counter Output:	-50.7 volts
Net Counter Output at Release:	+6 volts
Cathode-Follower Grid Voltage at Release, e_1:	73 volts (approx.)

One general characteristic of low-altitude bombing operation should be noted. Release always occurs when negative counter output due to speed largely cancels the positive output due to range. That is, the residual terms

due to S_o (and R_1 if used) are fairly small. Accurate sensitivity ratios are necessary for correct release, but the small bias-voltage difference and the load resistor across which counter outpout appears need be controlled to only moderate fractional accuracy.

Maximum usable sweep width is determined by linear modulation capability of the modulator and transmitter, while a minimum value is set by errors resulting from stray frequency modulation, due for example to vibration. Extra counts are produced by counter switching and by fixed error in the differentiated amplifier-output signal (see sections 2g and 3c of Chapter IV.) Minor variations of modulating frequency are compensated by the modulation wave-shaping method used (see section 4c of Chapter III.). It may also be noted that the *AN/APG-4* would have been more flexible for adaptation to other special uses if the entire bias-controlling resistor chain, and the entire sweep-control chain also, had been placed in the easily replaceable external altitude-compensation unit rather than in the main transmitter-receiver unit. Higher transmitter power might also have been valuable, and indeed was used in the experimental prototype units.

Calibration to make up for reasonable tolerances on component-part characteristics is obtained by adjustment of the altitude-compensating divider, as well as of modulation-amplifier gain and of bias applied to the counter load, using measuring methods described in Chapter VII. Residual range used in calibration of the bombing equipment must allow not only for range equivalent of the r-f lines but in addition for any distance by which the bombs may be mounted ahead of the antennas on the aircraft. Calibration of the altimeter used to control level bombing flight must be based on height of bombs above ground rather than on that of wheels above ground.

c. *Accessory Units SA-28/APG, RE-17/APG and C-141A/APG.* The necessity of flying at a definite predetermined altitude when using the standard *AN/APG-4* system is not attractive. A small servo unit, mounted in an A-N aircraft-instrument case and designated as *SA-28/APG,*[2] was therefore developed to compensate the system automatically for any altitude between 40 and 400 feet at which the bombing craft may happen

to fly. This accessory, used to replace the *SA-9A/APG-4* switch, contains a simple servo motor of the type shown in Fig. IV.-20, controlled by the limit-circuit relays of the altimeter. The output shaft of the servo, which mounts a linear follow-up potentiometer (r_2 of Fig. IV.-20), also drives three rheostats which control, respectively, the sweep width, speed-intercept and residual-range bias, and range-lead bias supply of the Sniffer. The two bias-control rheostats have non-linear resistance-rotation characteristics, while the required form of non-linear variation of sweep width with altitude is secured by using a linear rheostat shunted by a fixed resistor. The shunted rheostat is the r_1 of Fig. III.-12, with r_2 another suitably chosen fixed resistor; the switch *Sw* of the *RT-27/APG-4* is located in the circuit as shown in full lines in that figure.

The *SA-28/APG* altitude-compensation unit contains also the range-lead control and system power switch, as may be seen from Fig. VI.-10. The circuit of this unit is included in Fig. VI.-14. Altitude for which the Sniffer is compensated is continuously and automatically indicated

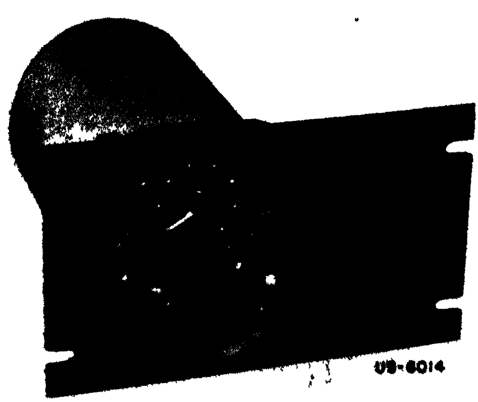


Fig. VI.-10. Automatic altitude-compensation unit for low-altitude bombing.

on the face of the unit. Two cables, one to the Sniffer transmitter-receiver unit *RT-27/APG-4* and one to the automatic-pilot receptacle of the altimeter transmitter-receiver unit **RT-7/APN-1*, are connected to receptacles at the rear of the compensation unit. Power for the motor

is derived from the altimeter, so that it operates even when the *AN/APG-4* is turned off. With this accessory, the pilot need only fly level at any desired altitude between 40 and 400 feet during the final stage of an attack, and need not seek a preset altitude; the bomb will still be released automatically at the range correct for his actual altitude and speed.

Connection of the *SA-28/APG* directly to the altimeter requires that the limit-relay contacts of the latter be rewired, and in such a way that three-light limit indication is no longer possible. To avoid this, a relay unit designated *RE-17/APG* was developed as a further accessory. This unit is connected by one cable to the auto-pilot receptacle of a normal altimeter, and by another cable to the altimeter receptacle of the *SA-28/APG* altitude-compensation unit. It contains two relays connected directly across the "too high" and "too low" lights of the normal limit indicator; these relays operate to drive the servo and its follow-up potentiometer in the directions respectively of lower and higher altitude. The *RE-17/APG* unit also contains a third relay, which transfers control of the altitude-limit counter from the *SA-28/APG* to the normal *SA-1/ARN-1* limit-setting switch of the altimeter whenever the *AN/APG-4* equipment is turned off.

Servo-motor power is supplied to the *SA-28/APG* unit from the *AN/APG-4* when the *RE-17/APG* unit is used, so the servo does not operate unless the bombing system is turned on. The *RE-17/APG* unit further contains calibrating rheostats for the servo follow-up circuit, so that the altimeter can be calibrated for proper bombing operation of the *SA-28/APG* compensation unit independently of the normal calibration of its limit-indicator circuits. When using the auxiliary relay unit, operation of altimeter limit lights is entirely normal except while altitude compensation of the *AN/APG-4* is actually required. Of course, while the servo is actually in use limit lights no longer indicate departure of the aircraft from a chosen altitude, but only the occurrence of corrections to the servo position as the aircraft changes altitude.

Some bomb-release mechanisms require more current than the contacts of the release relay of the *AN/APG-4* can

reliably supply. If the release relay operates repeatedly, as the target signal fades in and out after release, additional bombs may be dropped at incorrect ranges. Both of these troubles are avoided if the release relay of the Sniffer merely actuates a rugged booster relay, which upon operation locks itself in the operated position until manually reset, and which also closes a heavy bomb-release circuit. Such a latching booster relay is included in a small control unit, shown in Fig. VI.-11 and designated *C-141A/APG*. The control unit is intended to be mounted, with the *SA-28/APG*, in the panel plate shown in Fig. VI.-10.

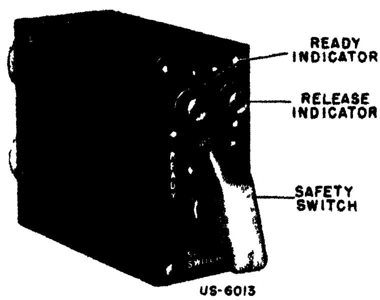


Fig. VI.-11. Control unit for radar bomb release.

This accessory unit also contains a safety switch, which in the "safe" position silences the *AN/APG-4* transmitter and removes power from the booster relay to prevent any possibility of bomb release by that relay. When turned on, this switch permits the transmitter to oscillate if the power switch of the *SA-9A/APG-4* or *SA-28/APG* is turned on; it also lights a red warning or "ready" lamp in the control unit, and enables the booster relay to operate if the release relay of the *AN/APG-4* closes even momentarily. When the booster operates, it actuates the bomb release and locks itself in the operated position; an amber "release" lamp in the control unit comes on with the release operation and remains on. Return of the switch of the *C-141A/APG* to safe position silences the transmitter, extinguishes both lamps, and returns the booster relay to its open condition, preparing the system for subsequent cycles of operation.

The complete automatic bombing system, with all accessories, consists then of two major and seven minor units.

The transmitter-receiver and two antennas of the **AN/APN-1* altimeter are required, and connect through relay unit *RE-17/APG* and altitude-compensation unit *SA-28/APG* to the transmitter-receiver unit of the *AN/APG-4* equipment. The latter is connected to its own two antennas and, through control unit *C-141A/APG,* to the bomb-release circuits of the aircraft. All these units, together with all interconnecting cables, weigh far less than a bombardier and serve to make low-altitude blind bombing with high accuracy a fairly easy matter for the pilot of even a fighter aircraft.

4. AZIMUTH CONTROLLING EQUIPMENT *AN/APG-6(XN)* FOR AUTOMATIC LOW ALTITUDE BOMBING

a. *Purpose and General Description.* The problem of supplying azimuth information in manned or unmanned aircraft, for the purpose of directing flight accurately toward an isolated surface target, has arisen in the use of f-m radar. Several experimental models of equipment to solve this problem, designated *AN/APG-6(XN)* and often called *Super-Sniffer,* were built. Successful flight tests of the experimental equipment had led to the construction of several pre-production samples of the final *AN/APG-6* design when changing tactical requirements caused the project to be dropped.

AN/APG-6 equipment performs the same low-altitude bombing function as does the *AN/APG-4.* It is able in addition, by use of a switched-lobe directive-antenna system, to close one or the other of two relays as the target deviates in azimuth to the right or left respectively of the direction established by the antenna orientation. The *AN/APG-6* and **AN/APN-1* equipments together, with their respective outputs connected to control an automatic pilot, make possible the automatic flight of aircraft so as to "home" at a chosen fixed altitude on an isolated surface target, and in addition to release a bomb at the proper range for a hit on the target. In this case, the *SA-1/ARN-1* altitude-limit switch will be used with the altimeter to set the altitude of automatic level flight, and the *SA-9A/APG-4* altitude-compensation switch will be used to set the *AN/APG-6* for automatic bomb release at the correct range for the chosen altitude.

Homing flight is not adequate for bombing in a cross wind, so the *AN/APG-6* equipment is actually used in conjunction with a gyroscopic stabilizer, which controls through a servo mechanism the orientation of the switched antennas. Provision is made in the stabilizer equipment to bring the aircraft automatically into a straight flight path which will directly intercept the target, as discussed in section 8 of Chapter V. The control output of the stabilizer may actuate the steering channel of an automatic pilot, or if manual flight control is desired the output may operate a Pilot Director Indicator (*PDI*). In the latter case, the pilot will fly level by use of the **AN/APN-1* indicator, and the automatic *SA-28/APG* altitude-compensation unit will be used to adjust the *AN/APG-6* for correct release at the actual flight altitude. Sequence of flight operations may be controlled by range-dependent relay action in the *AN/APG-6*.

The circuits in the *AN/APG-6* which utilize the audio beat-note output of the receiver for operating the bomb-release mechanism of the aircraft are similar to those of the *AN/APG-4* low-altitude bombing radar, as is the radio transmitter, which however has an increased r-f power output of two watts. The receiver is quite different, and operates on the side-band superheterodyne principle described in section 5b of Chapter III. The operating frequency of the *AN/APG-6(XN)* equipments was chosen as 515 megacycles per second, primarily because of the early availability of suitable Yagi antennas designed for the *ASB* search radar. Pre-production samples of the *AN/APG-6* were built to operate at 410 megacycles, using the later *AN/APG-4* antenna shown in Fig. III.-3.

A functional block diagram of the essential features of the *AN/APG-6(XN)* equipment is shown as Fig. VI.-12. The basic features of a side-band superheterodyne receiver have already been shown in Fig. III.-24. Essential differences between the receiver in that figure and the receiver of the *AN/APG-6(XN)* equipments are the use in the latter of a radio frequency of 515 megacycles per second instead of 1500 megacycles per second, and of an intermediate frequency of 30 megacycles per second instead of 120 megacycles per second. Minor differences are the

use, in place of the two crystal mixers of the earlier figure, of a 30-megacycle combined oscillator-mixer and of a balanced first detector, identical with that of the *AN/APG-4* equipment but tuned to 515 megacycles per second.

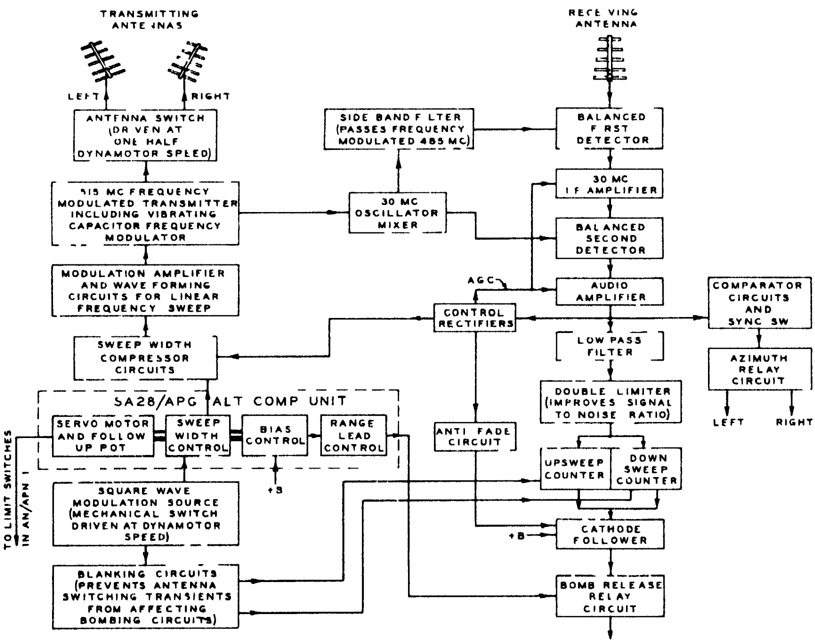


Fig. VI.-12. Block diagram of *AN/APG-6(XN)* azimuth-controlling radar for low-altitude bombing.

Operation of the side-band superheterodyne receiver is such that the frequency-modulated 515-megacycle received signal is heterodyned in the balanced first detector against a synchronously frequency-modulated 485-megacycle local mixing signal, to produce a 30-megacycle intermediate-frequency signal. The local mixing signal is the lower side band developed by mixing transmitter output with the output of a 30-megacycle local oscillator. Transmitter frequency modulation is removed from the received signal by heterodyning with the synchronously modulated local signal, so does not appear at the intermediate frequency. Low-frequency beat-note components appear at the output of the second detector, after the 30-megacycle intermediate-frequency radar signal has been heterodyned

against the 30-megacycle mixing signal from the local oscillator.

Transmitter-modulator unit, 30-megacycle oscillator-mixer, and side-band filter tuned to pass 485 megacycles per second ±½W are constructed as a single sub-assembly. Fig. VI.-13 shows the interior of the 25-tube *AN/APG-6(XN)* equipment and includes the transmitter, local oscillator

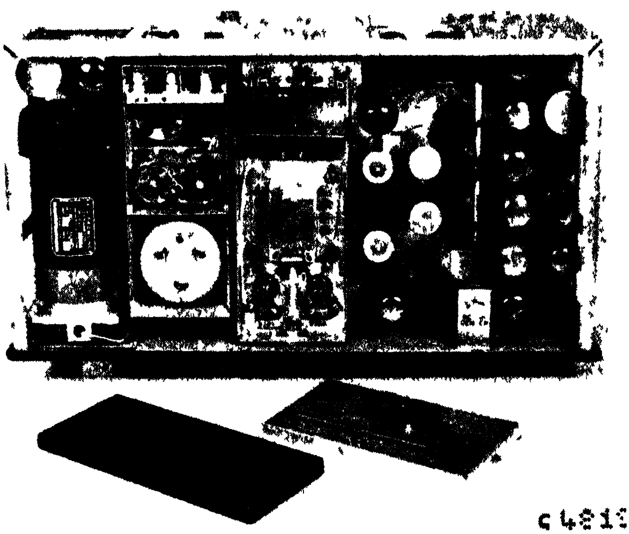


Fig. VI.-13. Interior arrangement of *AN/APG-6* (*XN*) equipment.

and side-band filter, all in the unit which is visible next to the dynamotor at the left of the figure. The vibrating modulator is at the end of this unit adjacent to the front panel and the side-band filter is at the rear end of the unit. The second detector of the *AN/APG-6(XN)* equipment is balanced against microphonics originating in the oscillator-mixer; it is mounted on the i-f and audio-amplifier sub-chassis seen at the right of the balanced first-detector unit in the figure. Transmitter, receiver and audio-amplifier circuits are shown schematically in the left half of Fig. VI.-14.

The low-frequency radar-beat signals from the second detector are amplified in a high-gain audio amplifier having a gain-frequency characteristic similar to that of

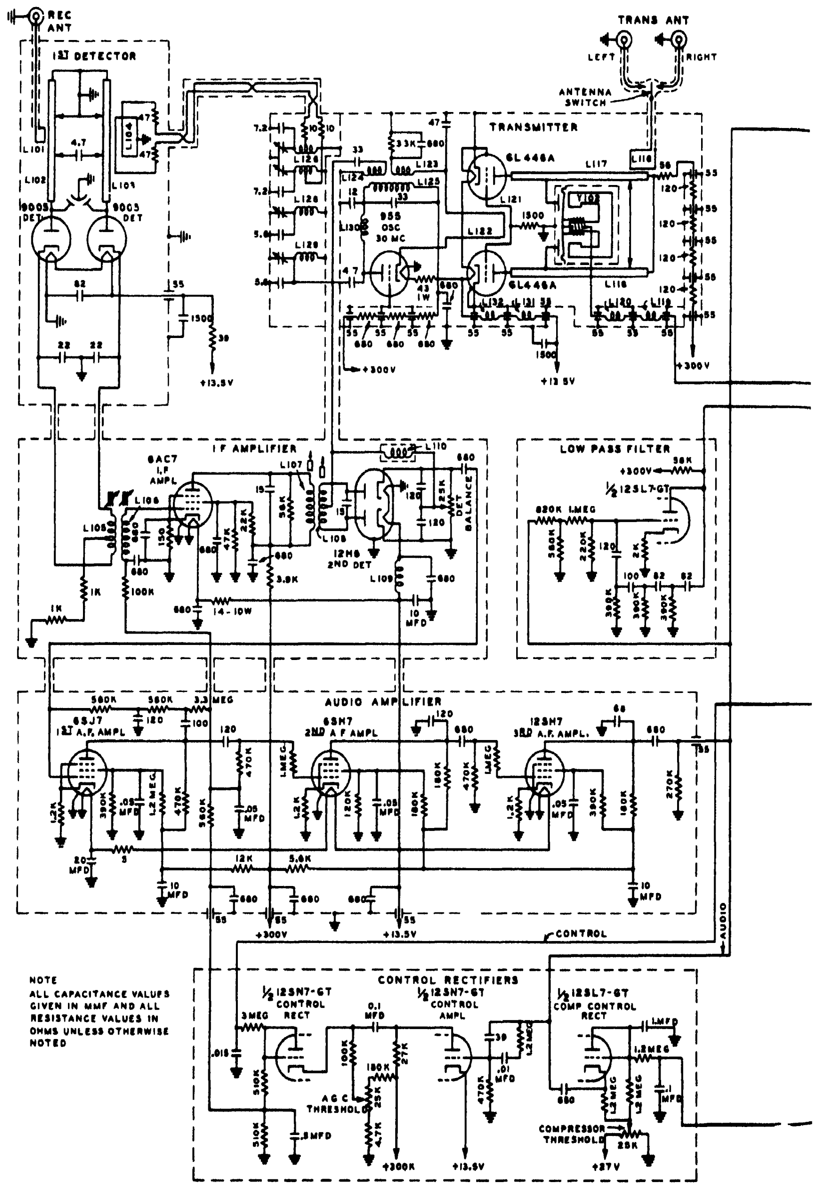
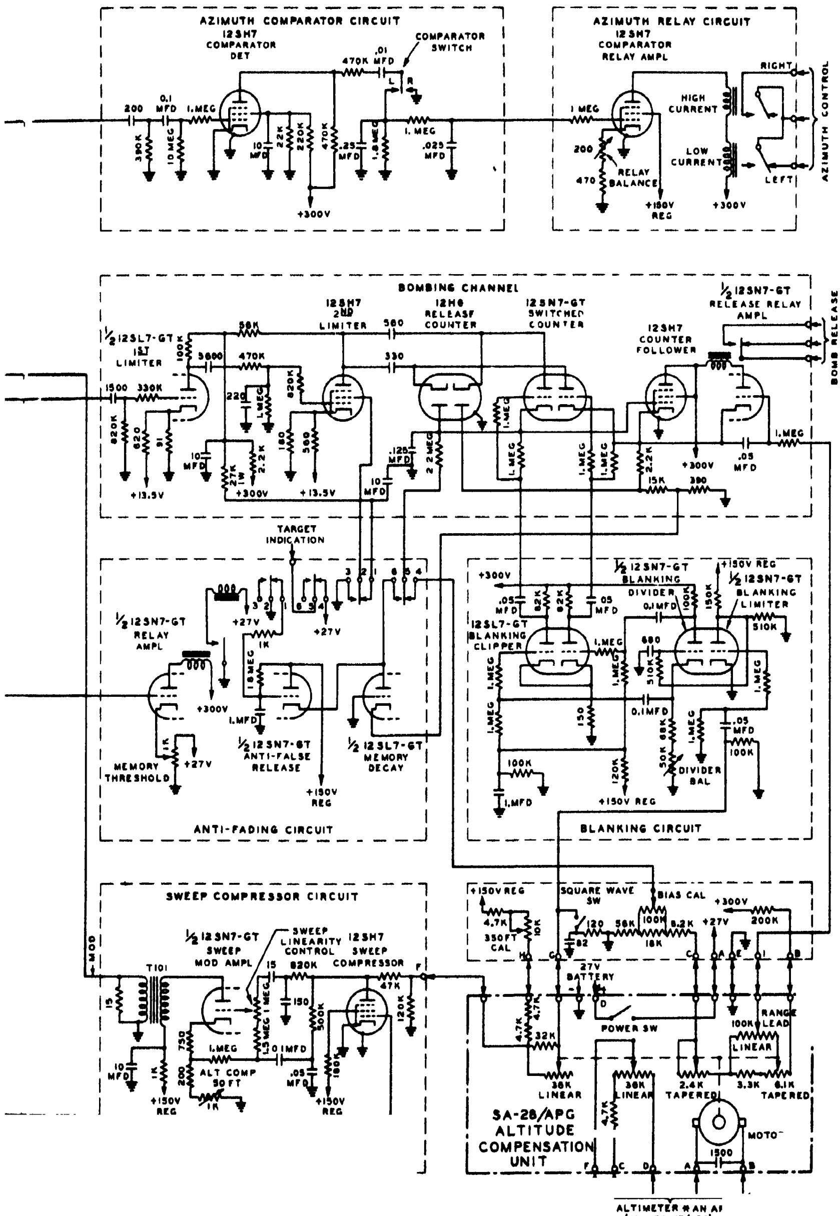


Fig. VI.-14. Functional circuit diagram of *AN/APG-6(XN)*



azimuth-controlling radar system for low-altitude bombing.

Fig. III.-28. Output from this amplifier feeds two separate channels through filters of different characteristics. One of these channels, fed through a low-pass filter to reduce disturbing noise, comprises limiters, counters, and relay circuits for utilizing range and speed information for bombing. This channel is similar to the corresponding portion of the *AN/APG-4*, with the exception that a counter "blanking" circuit is added to prevent unwanted transients from affecting the accuracy of operation of the counter circuits. Counter blanking has been described in Chapter IV., section 3e. A "memory" type of device to offset signal fading, described in section 2 of Chapter VII., is also added and operates as an accessory to the bomb-release circuits. The second channel, fed through a high-pass filter to reduce altitude signal while the aircraft is banked in turns, comprises the circuits necessary to convert signal-strength variations determined by target azimuth, which are produced by antenna switching, into useful operation of a pair of relays. This azimuth-determining channel will be described later. Bombing and azimuth-determining circuits are shown schematically in the right half of Fig. VI.-14, as are the modulating circuits and the *SA-28/APG* altitude-compensation circuits.

The transmitting-antenna system comprises two Yagi antenna arrays, mounted with a fixed angular displacement of 30 degrees between the center lines of their respective horizontally directed beams. For gyro-stabilized orientation, a remotely controlled servo unit of sufficient power to rotate the antennas is coupled to the antenna pair; this permits the mean azimuth of transmission to be displaced as much as 30 degrees from the heading of the aircraft. This servo and the stabilizer which controls it are described in detail in the Handbook of Maintenance Instructions for Model *MX-247/APG(XN)* Stabilizer. A general description of the way in which the stabilizer acts to provide flight-control data will be given later. Use of a remotely operated servo unit to position the transmitting-antenna pair allows placement of the antennas on the aircraft to be determined primarily by electrical considerations affecting the operation of the equipment, while the gyro stabilizer is mounted in any convenient location.

A separate 50-ohm coaxial transmission line is connected between each antenna and one contact of a single-pole, double-throw switch having its blade fed by the transmitter. This switch is operated by a cam driven from a two-to-one reduction gear on the dynamotor shaft; it is located in the main transmitter-receiver unit of the equipment. Transmitter power is thus fed to the two antennas alternately through the switch, at a frequency of alternation equal to one half the rotational speed of the dynamotor. The housing of the antenna-selecting switch may be seen at the front end of the dynamotor in Fig. VI.-13, with the antenna-cable receptacles projecting through the front panel of the equipment. A switch which must operate synchronously with the antenna switch is used in the azimuth-determining circuit and is also contained in this housing. Double extensions on the dynamotor shaft permit mounting the mechanical switch which serves as a modulation-voltage source on the opposite end of the dynamotor frame from the gear-driven antenna switch. The modulating switch is identical to that used in the *AN/APG-4* equipment and operates at dynamotor speed. It is evident that one complete frequency-modulation cycle of the transmitter occurs for each position of the antenna switch.

Actually, the antenna switch is somewhat more complicated than the simple single-pole, double-throw variety; it is arranged to have "make before break" contacts so timed that for about one per cent of the switching cycle both antennas are connected to the transmitter. The rest of the cycle is divided equally between the two antennas acting singly. The modulation switch is phased mechanically with respect to the antenna switch so that one peak of each cycle of frequency modulation occurs at the center of each antenna-switch overlap interval. Antenna switching produces load changes on the transmitter, because it is not practicable to build a switch in which there is consistently neither dead space nor overlapping. Rapid changes in transmitter loading such as occur due to switching action produce undesired transient signals, as previously mentioned. Presence of a definite overlap time holds these transients to a practical minimum. Such transients have no adverse effect on the azimuth-determing circuits, but would be definitely detrimental to the

operation of the bombing circuits if the switched counters were not blanked.

Reflected signal from the target is picked up on a single fixed Yagi receiving antenna, mounted so that the center line of its single major directive lobe is parallel to the longitudinal axis of the aircraft. Thus, reflected signal is received alternately from one transmitting antenna or the other, except for a short interval of time (antenna-switch overlap time) when reflected signal from both transmitting antennas is received. Fig. VI.-15 shows all three antenna directive patterns as dashed curves, under the condition that the receiving-antenna orientation is centered with respect to the two transmitting antennas.

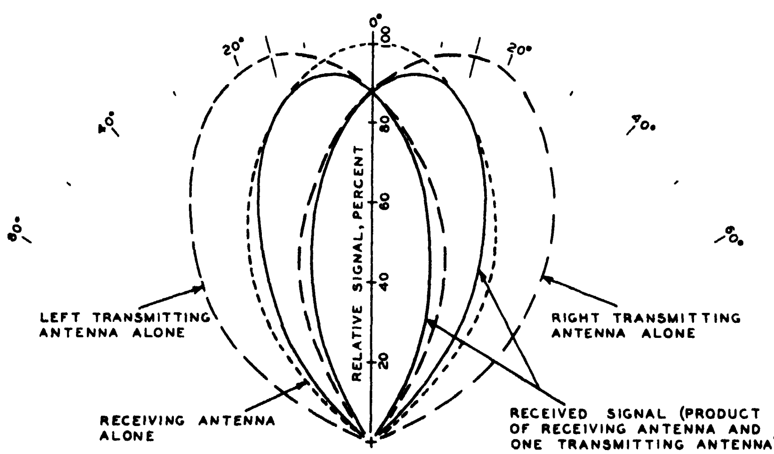


Fig. VI.-15. Horizontal directive patterns of *AN/APG-6* antennas.

Each individual antenna has the same directive pattern in both *AN/APG-4* and *AN/APG-6* systems.

The products of transmitting and receiving field patterns, represented by the full-line curves of Fig. VI.-15, show that a reflecting target on the center line would give equal received-signal strength for transmission from either of the switched antennas. Should the reflecting target be displaced to the left of the center line, received signal would be greater when the left transmitting antenna is energized and less when the right transmitting antenna is energized. The antenna patterns indicate development of

a total difference in received-signal strength under the two transmitting conditions of about 3 per cent per degree of azimuth deviation of the target from the line of equal signal strength. Azimuth-determining circuits provide means for detecting and utilizing this signal-strength differential to operate right-left relays. Operation of these relays may be used to guide the aircraft or to cause a servo system to rotate the transmitting-antenna pair in the proper direction to reduce the signal-strength differential to zero.

b. *Azimuth Determining Circuits.* Operation of the azimuth-determining or "comparator" circuits of the *AN/APG-6* is rather different from that of the various system elements previously covered, so will be separately described here. The arrangement used is noteworthy for the sensitivity attained by relatively simple means.

As has been explained already, antenna switching produces square-wave amplitude modulation of the radar beat-note signal, dependent in degree and polarity on the deviation of target azimuth from the line of symmetry of the switched-antenna directive patterns. So far as azimuth determination is concerned, the radar beat serves merely as a carrier signal for this amplitude modulation. Frequency modulation of the transmitter must be so controlled that this carrier will be passed by the audio amplifier for any range at which azimuth determination is required. A "sweep compression" circuit is necessary to accomplish such a result for the longest ranges, and will be described later in this section.

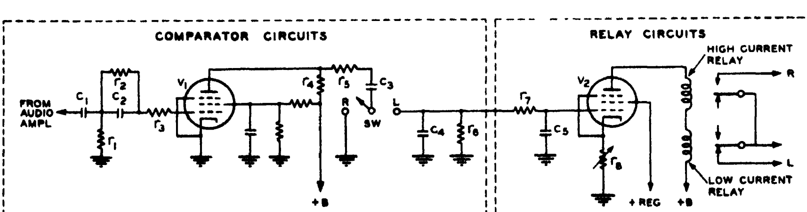


Fig. VI.-16. Circuits for determining target azimuth.

Fig. VI.-16 shows the circuits of the azimuth-determining comparator and its associated relay amplifier. Low frequencies in the beat-note carrier signal are greatly

reduced by using a small capacitance in the input-coupling circuit C_1, r_1. At short ranges, where the range-beat signal from the desired target is strong and of low frequency, this does no harm. At long ranges, where the desired-target signal is weak and of high frequency, low-frequency sea-return signal received in banked turns is prevented from acting as a strong target off to one side.

The comparator tube V_1 itself operates as a grid-leak detector. Grid current through leak r_2 biases the tube well beyond cut off, in the presence of high-level audio input. The long time constant of the grid-leak and condenser combination r_2, C_2 maintains this bias steadily at such a value that the peaks of the azimuth-modulated radar-beat signal just drive the grid to zero. The plate of the comparator tube therefore swings from supply voltage to practically zero voltage on a portion of each beat-note cycle, while the antenna giving the stronger signal is connected. Since the grid bias can not decay appreciably within an antenna-switching cycle, grid voltage will not swing up to zero nor plate voltage down to zero on beat-signal cycles produced while the weaker antenna is active. Series resistor r_3 prevents extremely rapid charging of C_2, and so prevents undue alteration of bias by strong but very brief transient-signal peaks which may be produced by antenna switching or other disturbances.

Average voltage at the plate of V_1 will be below the supply level by an amount depending on the duration and strength of the plate-current pulses, at radar-beat frequency, through plate load r_4. Duration of these pulses will depend on the amplitude of the strongest beat-note signal applied to the grid circuit, and their momentary strength on the degree to which the momentary amplitude of the input approaches its bias-setting strongest value. Average drop of plate voltage below supply level E_0 while the right antenna is active may be called e_R, and average drop for left antenna, e_L. A typical strong-signal drop for the stronger antenna is 55 volts, with a 3-per-cent smaller signal from the weaker antenna then giving an average drop of 50 volts.

The integrating or averaging connection of the transfer circuit r_5, C_3 insures that the voltage at the high

terminal of C_3 will be substantially the average plate voltage of V_1. This is either $E_o - e_R$ or $E_o - e_L$, depending upon the antenna in use. Comparator switch *Sw* transfers the low terminal of C_3, synchronously with the antenna switching, from connection to ground while the rightward-pointing antenna is active to connection to the output-load circuit r_6, C_4 while the left antenna is active. The voltage across transfer capacitor C_3 is therefore $E_o - e_R$ in the former and $E_o - e_L - e$ in the latter case, where e is voltage across the output circuit. If these voltages are different, an increment of charge will enter C_3 each time the larger voltage is applied. With antenna-switching frequency f_a, the average current transferred through C_3 by this sequence of charge increments is

$$i = [(E_o - e_L - e) - (E_o - e_R)] f_a C_3. \tag{VI.3}$$

For a steady-state condition, output voltage e will take such a value that the current e/r_6 leaking through r_6 is just able to consume during each switching cycle the charge increment transferred by C_3. That is, equating currents,

$$e = (e_R - e_L) f_a C_3 r_6 / (1 - f_a C_3 r_6). \tag{VI.4}$$

Comparator-output voltage e is thus proportional to the difference between the signal levels associated with the two antennas, while the polarity of e depends on which signal is the stronger. So long as C_3 can be fully charged through r_5 while the switch is in one position, the value of r_5 is immaterial except as a smoothing element. The time constant with which output voltage e can follow changes in $e_R - e_L$, or target azimuth, is controlled by the value of C_4 in conjunction with that of r_6 in parallel with an equivalent resistor $1/(f_a C_3)$.

Comparator output will not be disturbed by antenna-switching transients if the circuit of C_3 is open during these transients. This condition is attained by allowing the switch *Sw* to open from one position before antenna switching starts on each half of the switching cycle, and not allowing *Sw* to close in the other position until after antenna switching is complete. The time sequence of switch operation used is shown in Fig. VI.-17.

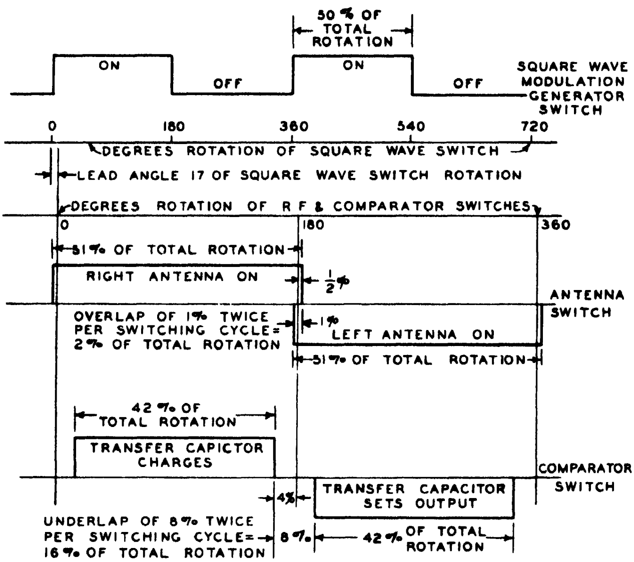


Fig. VI.-17. Timing cycle and phase relation of modulating, antenna, and comparator switches.

In the *AN/APG-6*, the comparator-output signal is applied through a ripple filter r_7, C_6 to the grid of a current amplifier V_2, which in turn has a high-current relay and a low-current relay in series in its plate circuit. Cathode resistor r_8 is adjusted so that the positive comparator-output voltage required to actuate the high-current relay is equal to the negative comparator output required to open the low-current relay. This balance is stabilized by feeding the amplifier screen from a regulated supply voltage. In the normal condition, with zero comparator output, the low-current relay is held actuated and the high-current relay remains open. Relay operation is arranged to occur for an amplifier input of ±2 volts, corresponding to a target-azimuth deviation of ±½ degree for strong radar signals or ±1 degree for signals 3 decibels above noise. Greater azimuth sensitivity is easily obtainable if needed.

The simplicity of obtaining control action with f-m radar equipment is again emphasized by this arrangement. Aside from the additional antenna required to provide more data, target-azimuth determination requires only addition

to the equipment of two synchronously driven two-way selector switches and one tube with simple associated circuits. To provide control actuation on the basis of azimuth, one more tube and two relays of simple form are required.

Consideration of effective working ranges of the low-altitude bombing and azimuth-determining functions of the equipment shows that while relatively short ranges are sufficient for bombing, ranges several times as great are required for establishing the aircraft on the most direct course to intercept the target. For the maximum design altitude of 400 feet and maximum closing rate of 350 knots, the range from target at release will be approximately 3000 feet and at that point the aircraft must already be on an interception course.

Factors determining the range at which usable signals are needed for azimuth corrections prior to release are: stability of sighting platform, resolution of the sight, rate at which steering corrections can be made, and manner in which steering is controlled by target azimuth. Tests on *AN/APG-6(XN)* equipment as used in an *SNB* aircraft have indicated that usable signal for azimuth corrections needs to be received three to four miles from the target in order to assure that the aircraft will be accurately established on a direct interception course by the time that it reaches the bomb-release point.

Required transmitter sweep widths for bombing vary between 4.1 megacycles per second at an altitude of 50 feet and 1.7 megacycles per second at 400 feet altitude. For release at maximum altitude and maximum closing speed, the sum of range-beat frequency and speed-beat frequency is about 3000 cycles per second, so a receiver band width of 4000 cycles per second is ample for bombing. However, using a minimum sweep width of 1.7 megacycles per second at a range of 4 miles results in a range frequency of 15000 cycles per second, and for maximum sweep of 4 megacycles the range frequency at four miles is about 35000 cycles. Both of these range frequencies are well outside the audio band required for bombing. Severe reduction of signal-to-noise ratio for bombing would result if a wide-band beat-frequency amplifier were used to admit such high

beat frequencies at maximum azimuth-determining range.

An obvious solution is to reduce the sweep width to values below the bombing minimum when the aircraft is at greater ranges than maximum bombing range. Minimum attainable sweep width is restricted by residual frequency modulation of the transmitter, due in particular to microphonics in the vibrating-capacitor modulator. Residual sweep in *AN/APG-6(XN)* equipment is of the order of 0.3 megacycle per second. The audio-amplifier frequency characteristic, which has the general form shown in Fig. III.-28 except for a steeper rise in the operating region, is peaked at 6000 cycles, corresponding for a range of three miles to a sweep width of approximately 0.85 megacycles per second. No further reduction in sweep width is required for ranges somewhat greater than three miles, since the amplifier does not cut off very sharply on the high-frequency side of its pass band. Reduction of higher-frequency response for the bombing channel is required to reduce noise and possible interference from distant targets. This reduction is accomplished by means of a tube with selective negative feed back applied to it by a resistance-capacitance network. The high-cutting tube is interposed between the output of the audio amplifier and the grid of the first limiter tube of the bombing channel, reducing the frequency of overall peak response in that channel to 4000 cycles or less.

Automatic means is provided in the equipment to vary sweep width from a minimum value at maximum range to the value necessary for functioning of the bombing circuits. This automatic sweep-width compression circuit operates by controlling, in accordance with received-signal strength, the amplitude of the square-wave voltage applied to the modulator wave-forming circuit. When the received radar-beat frequency has fallen to the order of the peak response frequency of the bombing channel, the sweep-width control tube will normally be biased beyond cut off. Sweep width will then be controlled as a function of altitude by the *SA-28/APG* automatic altitude-compensation unit alone.

Fig. VI.-18 is a schematic diagram of the sweep-width compression circuit. Tube V_1 of this figure is a diode control rectifier fed from the output of the audio-frequency

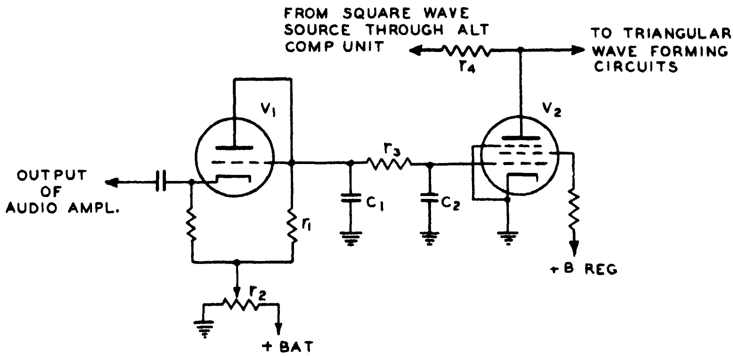


Fig. VI.-18. Automatic sweep-compression circuit.

amplifier. Rectification of the audio signal produces across the diode-output resistor r_1 a direct negative control voltage proportional to signal strength. Introduction of a controllable positive bias voltage into the diode circuit from voltage divider r_2 provides a signal-level threshold, by ensuring that the grid of sweep-compressing tube V_2 is held at cathode voltage in the absence of rectified output across r_1. Capacitors C_1 and C_2 serve with resistor r_3 as a smoothing filter to prevent audio frequencies from appearing on the grid of V_2.

In the absence of sufficient signal to over-ride the delay or threshold voltage introduced by r_2, V_2 is zero biased and its plate-to-cathode resistance is approximately 10,000 ohms. Square-wave plate voltage for V_2 is supplied by the *SA-28/APG* automatic altitude-compensation unit, through r_4. The combination of r_4 and the resistance of compressor tube V_2 to ground serves as a voltage divider interposed between the altitude-compensation unit and the modulator wave-forming circuits. Increasing audio-amplifier output produces a negative control voltage across r_1. When this exceeds the positive threshold voltage from r_2, it causes the grid of V_2 to go negative. This in turn causes the plate-to-ground resistance of V_2 to increase with increasing signal level.

When finally the rectified signal is sufficient to apply cut-off bias to the grid of V_2, conduction in this tube becomes negligible and it no longer acts in conjunction with r_4 as a voltage divider. Under this condition, modu-

lation voltage and therefore sweep width is determined only by the automatic altitude-compensation unit. Sweep-width compression is then no longer required, because the bias on V_2 is produced by the passage of strong beat signal through the amplifier. The purpose of compression was merely to insure the presence of such signal to act as a carrier for the azimuth data.

Operation of the sweep-width compressor circuit is limited in practice to the negative-slope or high-frequency region of the audio-frequency amplifier-gain characteristic. Increases in received signal due to rising amplifier gain, as range and range-beat frequency decrease, permit the compression circuit to increase sweep width. This in turn tends to increase range-beat frequency, since changes in sweep width result in changes in range frequency for a given range. The compression circuit thus acts to maintain the received frequency at or near the peak of the audio-amplifier characteristic as range decreases. The result is effectively an automatic frequency control which adjusts sweep width to maintain received frequency within the pass band of the audio amplifier. Means for automatically varying sweep width are not necessarily limited to control by signal strength, however. Control may for example be developed as a function of range or time from target.

Because azimuth sensing in the comparator circuits is accomplished by comparing signal amplitudes, automatic gain control to prevent limiting action in the audio amplifier is essential. Control voltage for automatic gain control is derived from a high-level control rectifier, also used to actuate the anti-fading accessory or "short-time memory" circuits, which are included in Fig. VI.-14 and discussed separately in section 2 of Chapter VII. It is evident that, if a-g-c acts to limit the amplifier output to a level below that required to assure complete cut off of the sweep-width compression tube, inaccuracies in bombing will result because of insufficient sweep width. Satisfactory operation of the sweep-width compression circuit is therefore dependent upon proper relative settings of sweep-width-compression threshold control, r_2 of Fig. VI.-18, and of automatic-gain-control threshold.

c. *Control of Aircraft Steering*. Operation of the azimuth relays of the *AN/APG-6* furnishes the necessary information to steer an aircraft, either manually or automatically, so as to keep it pointing directly at the radar target. This type of steering will not cause a bomb released from the aircraft to strike the target, however. To bomb, it is necessary to establish the aircraft on a radial or direct-interception course, for which there is no relative motion of aircraft and target transverse to the line joining them.

There is a straightforward navigational procedure, described in section 8 of Chapter V., that will bring an aircraft smoothly into an interception course. This procedure requires a sighting means, represented by the switched antennas and azimuth-determining channel of the *AN/APG-6*, which is rotatable on the aircraft and is kept trained on the target. It also requires a steady direction reference, whichmay be provided by a gyroscopic stabilizer.

The fundamental element of the stabilizer is a gyroscope wheel kept spinning about a nominally horizontal axis. The gyro, with the ring in which its bearings are mounted, is free to tilt about a horizontal axis perpendicular to the spin axis. This tiltable assembly, with the tilt-axis bearings carrying it mounted in a nominally vertical ring called the *cardan* ring, is free to turn about a vertical cardan axis with bearings fixed to the aircraft. It is the basic property of the gyroscope when undisturbed that the direction of its spin axis will remain fixed in space, however the aircraft attitude may change. It is a further basic property that when disturbed by a torque applied about the cardan axis the gyroscope will not turn about that axis, but will instead rotate about the tilt axis or *precess*. A torque about the tilt axis will similarly cause precession about the cardan axis.

The gyroscope is used in the stabilizer to adjust potentiometers mounted on its cardan axis. These potentiometers, as well as the cardan bearings, produce a torque about that axis whenever the aircraft yaws with respect to the fixedly directed gyroscope. Torque to operate the control potentiometers as the aircraft yaws is the useful output of the stabilizer unit, but must be obtained without

disturbing the fixity of the reference direction in azimuth established by the gyroscope spin axis. A servo mechanism is necessary to reconcile these two requirements. Cardan-axis torque causes the gyro to precess about the tilt axis, the direction of tilt depending upon the direction of the torque. This tilting closes one or the other of two control contacts, and thus activates a servo motor which turns the gyro mechanism with respect to the aircraft so as to maintain the gyro axis fixed as the aircraft yaws. The servo supplies the torque necessary to adjust the potentiometers and to overcome bearing friction, thus arresting and ultimately cancelling the tilt precession that initiated servo action.

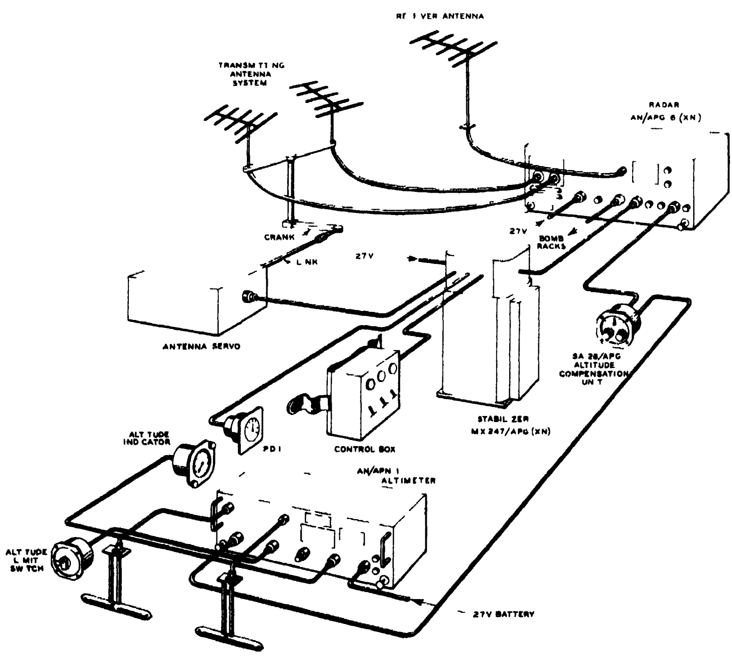


Fig. VI.-19. Complete system for guidance of aircraft in automatic low-altitude bombing.

Fig. VI.-19 is a pictorial diagram of the complete system required for guiding an aircraft in altitude and azimuth toward a surface target and releasing a bomb automatically at proper range. The **AN/APN-1* altimeter shown has already been described in section 2 of this chapter, and its use to compensate the bomb-release computer through

the *SA-28/APG* unit has been described in section 3c. Additional equipment required to utilize the azimuth data from the *AN/APG-6* radar comprises four units. These are the *MX-247/APG* gyro stabilizer proper, antenna servo, control box, and Pilot-Director Indicator (*PDI*). In the actual *MX-247/APG(XN)* equipment, a separate junction box was also used, but such a fifth auxiliary unit is not essential.

Fig. VI.-20 is a block schematic diagram showing essential features of the stabilizer, antenna servo, and steering control. As indicated, steering may be done either by a

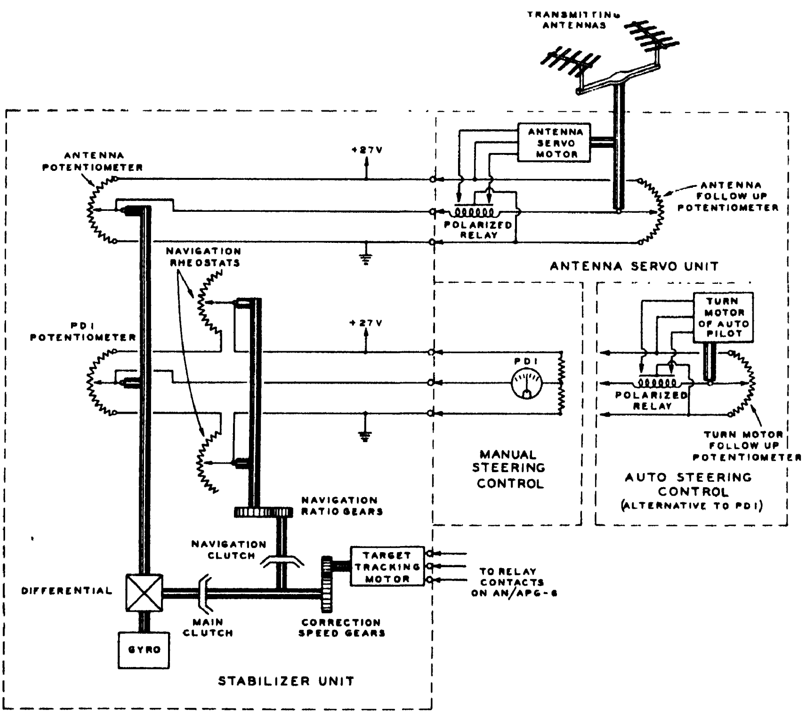


Fig. VI.-20. Block schematic diagram of stabilized steering system.

human pilot under *PDI* guidance or directly by controlling an automatic pilot. The small control box shown in the pictorial diagram, Fig. VI.-19, contains a main stabilizer and gyro power switch, as well as the necessary switches to operate magnetically the main clutch and the navigation clutch shown in Fig. VI.-20. When the main clutch is engaged but the target-tracking motor is stopped, yawing

of the aircraft results in causing the resistive elements of both the antenna and the *PDI* potentiometers of the stabilizer, which are attached to the airplane, to rotate with respect to their brushes, which are geared to the gyro shaft and consequently fixed in azimuth. One of these potentiometers is connected in a bridge circuit with the follow-up potentiometer of the antenna servo. This bridge circuit will be unbalanced by motion of the stabilizer, causing current to flow through a sensitive polarized relay connected to the bridge output. This relay in turn activates the antenna-servo motor, which rotates the antennas to return the bridge to a balanced condition. Thus the antennas are kept aligned in azimuth with the stabilizing gyro while the aircraft turns with respect to both, for any rate of turn up to 15 degrees per second.

Motion of the other cardan potentiometer and consequent current unbalance in the *PDI* bridge also results from yaw; this causes a deflection of the *PDI* proportional to the angular displacement between aircraft and gyro. By an alternative arrangement, similar to the antenna servo and connected as indicated in place of the *PDI*, steering of the aircraft may be done automatically through the auto pilot. Operation of a target-tracking servo controlled from the azimuth relays of the *AN/APG-6* and acting on the potentiometer shaft through differential gearing results in superimposing target-azimuth changes upon the gyro control of the antenna and *PDI* potentiometer arms. Thus the antennas may be made to rotate, with respect to the gyro-stable reference, so as always to point directly at the radar target. Indications given by the *PDI* are modified in accordance with this antenna rotation.

The above method of utilizing the azimuth information from the *AN/APG-6* leads to a homing course, since it causes the *PDI* to indicate in proportion to the angular displacement between the antennas and the aircraft heading; such a course is useless for bombing. Direction indications useful in establishing a *navigation* or direct-interception course are obtained by engaging the navigation clutch as well as the main clutch. This results in driving a pair of navigation rheostats in the *PDI* circuit, at a separately determined rate, from the target-tracking servo which also

drives the antenna and *PDI* potentiometers. Target-azimuth information supplied by the *AN/APG-6* equipment now causes the balance point of the *PDI* bridge to occur with the *PDI* potentiometer off its central position, because of the displacement of the navigation rheostats. That is, a centered *PDI* indication is obtained with the aircraft heading turned somewhat away from the direct line of sight of the target.

For example, changing target azimuth may require the antennas to be rotated one degree to the left with respect to the gyro, in order to keep them trained on the target and so balance the comparator circuit in the *AN/APG-6*. In a homing approach, the *PDI* would call for a one-degree left turn and the aircraft would be so turned to keep it headed directly toward the target. By use of the navigation rheostats, driven through a suitably chosen gear ratio, the *PDI* may be made to call instead for perhaps a 3-degree turn to the left. Use of the navigation drive and rheostats thus permits the heading of the aircraft to be turned away from the line of sight to the target in orderly fashion. The aircraft is thereby made to follow the type of track illustrated roughly in Fig. V.-13, and so to establish itself in a direct-interception approach.

If the aircraft yaws momentarily, perhaps because of rough air, the stabilizer and antenna servo act rapidly to keep the antennas trained on the target and the *AN/APG-6* and target-tracking servo are not affected. The *PDI* then indicates directly the yaw to be corrected, without alteration by the navigation rheostats. *PDI* azimuth-deviation indications are enlarged by a "navigation ratio" greater than unity only with respect to such antenna rotations in space as are caused by actual slow drift of the target across the line of sight. The indications thus increased serve to turn the aircraft so as to reduce the rate of crosswise target drift.

Operation of *AN/APG-6* equipment with gyro stabilizer requires that limits be placed on the angular displacement between antennas and gyro-stabilized reference axis, because of the limited angular-motion capability of the potentiometers driven by the stabilizer. In order to utilize most effectively the permissible angular displacement

of the antennas, limited in the actual equipment to plus or minus 30 degrees, it is necessary that they be centered, or aligned parallel to the longitudinal axis of the aircraft, until a target run is started. The navigation rheostats should also be brought before the start of a run to the position for which the *PDI* bridge circuit is balanced when the antenna potentiometer is centered.

Control-contact segments, having an open-circuit position aligned mechanically with the longitudinal axis of the aircraft, are associated with both the antenna potentiometer and the navigation rheostats. The target-tracking servo is controlled by the *AN/APG-6* azimuth relays only when a master switch on the system control unit of Fig. VI.-19 is thrown to call for Sniffer operation. At all other times while the stabilizer is operating, the target-tracking servo is under control of the centering segments.

Electrical interlocks in the centering circuit assure that the target-tracking servo is first controlled by the navigation-rheostat centering segments. When centering of these rheostats is completed, the navigation clutch is disengaged automatically and the servo is placed under control of the antenna-potentiometer centering segments to center this potentiometer, and with it the antennas and the *PDI* potentiometer. When all centering is complete, the main clutch connecting the target-tracking servo is automatically disengaged and the shaft to the antenna and *PDI* potentiometers is locked to the aircraft structure by a mechanical latch. Under this condition no mechanical caging or restraint of the gyro is necessary; the gyro-servo motor, controlled from the precession-axis contact segments of the gyro, will compensate for torque caused by friction in cardan-axis bearings and differential gears and so will maintain the gyro spin axis horizontal as the aircraft turns. Since the control potentiometers of the stabilizer are always centered when Super-Sniffer operation is started by throwing the master switch on the control unit, and so applying gyro control by engaging the main clutch, initial orientation of the gyro spin does not matter. Only the directional stability of the gyroscope is important.

Automatic flight is readily achieved by using the alternative circuit of Fig. VI.-20, in which the *PDI* is replaced by a polarized relay controlling the turn motor of an automatic pilot. The follow-up potentiometer shown driven by the turn motor serves to rebalance the steering bridge with the turn motor displaced from its neutral position. Displacement of the turn motor is thus made proportional to the departure of aircraft heading from a value which is established by the line of sight to the target, in conjunction with the drift correction introduced by the navigation rheostats. Since rate of turn of the aircraft is proportional to the turn-motor position, there results a rate of turn proportional to heading change required and directed to reduce that change. This is the condition for stable turning motion of the aircraft, with simulated viscous damping.

In general, the principles discussed in section 2c of this chapter, on automatic flight control by radar altimeter, govern stability when utilizing radar azimuth information either for *PDI*-manual or for automatic steering of an aircraft. Optimum navigation ratio is determined by the stability of the sighting platform (in yaw this platform is the gyro, but in roll and pitch it is the aircraft) and by azimuth resolution of the antenna system. Speed at which course corrections may be made is determined by the rapidity with which the aircraft will follow such corrections. Aircraft may be subject to wild yaw oscillations about the proper heading if speeds of response of aircraft, automatic pilot and azimuth-determining means are not correctly related. This may be overcome in some cases by limiting the speed of the target-tracking servo.

At maximum range, differences between a homing course and a navigation or interception course are slight but target resolution is poor; satisfactory operation results only if a simple homing course is used during the initial phase of the approach. Where difference between homing and navigation courses becomes substantial, at perhaps one to one and a half miles from the target but in any case at a range depending on the relative transverse speed of the target, the navigation clutch should be engaged. This may be done automatically by the target-indicating circuit

of the *AN/APG-6*, on the basis of signal strength above a preset threshold. Target indication is provided by the anti-fading accessory which is discussed in section 2 of Chapter VII.

As the aircraft approaches the target, the angle subtended by the target becomes greater and, as in the case of a broadside attack on a ship, multiple reflections from various parts of a complex target will cause sudden, random course corrections to be applied. Such a condition results in violent heading changes if navigation corrections are being applied at short ranges. The navigation clutch may be disengaged automatically, a few seconds before bomb-release range is reached, by control signal from a release-warning accessory circuit, such as one of those described in section 1 of Chapter VII., applied to the *AN/APG-6*. [This circuit was not regularly included in the *AN/APG-6(XN)*]. After the navigation clutch has been disengaged, the aircraft may be flown in to the bomb-release distance on a homing basis, retaining the drift correction already established. Alternatively, the target-tracking servo may be stopped when the navigation clutch is disengaged by the release-warning signal. The aircraft will then be steered in to release by the gyro stabilizer only, on the heading last set up by radar before disengagement.

Other methods of arriving at the final navigation course are also possible. For example, a homing course may be flown between predetermined range limits with the navigation clutch engaged, but with the navigation rheostats electrically disconnected from the *PDI* bridge circuit, thus mechanically storing navigation information. This information may finally be applied to the aircraft course as one major drift correction, based on all accumulated data. Navigation ratio may if desired be made a function of range or time from target, thus providing optimum ratio at all times. As indicated by the operating sequence described, it seems desirable to use unity navigation ratio (homing course) at maximum range, to increase this ratio to the maximum stable value at medium ranges, and to reduce it again to unity, or even to zero (fixed-heading gyro steering), just before bomb release.

d. *Operation.* Aircraft utilizing *AN/APG-6* equipment would be directed to the target area in tactical operations by search radar or other suitable equipment. Such equipment need provide only moderately accurate range and azimuth information. Initiation of operation upon arrival in the target area requires that the aircraft be flown at a suitable low altitude and headed essentially toward the target. The master switch on the stabilizer-system control unit must be thrown, at a range depending on target size but approximately four miles, to transfer control of antenna orientation and *PDI* indication from the centering segments of the stabilizer to the *AN/APG-6*.

This initial procedure results in *PDI* operation to produce a homing course. Operation may later be changed as range decreases, so as to develop an interception course, by energizing either manually or automatically the navigation clutch of the stabilizer. Altitude must be held between 40 and 400 feet during the bombing portion of the approach only, although excessive altitude earlier may result in loss of signal. If the aircraft is flown to hold the *PDI* centered, it will be brought in on a proper course and a bomb will be released at a proper range to secure impact on the target. Once the approach has been turned over to the Super Sniffer, the pilot need only maintain some constant altitude within the permitted range and keep the *PDI* centered; all operation is then automatic. Upon completion of the run, control of antennas and *PDI* is returned to the centering segments of the stabilizer.

When *AN/APG-6* equipment is used in pilotless aircraft, transfer of control of the target-tracking servo from centering contacts to *AN/APG-6* relays, as well as selection of altitude at which the **AN/APN-1* altimeter holds the aircraft under automatic control, is accomplished by remote radio-control link.

Flight tests of the *AN/APG-6(XN)* equipment have been concerned primarily with its azimuth-controlling function. In very limited bombing tests, however, there was some indication that imperfect stability of counter blanking as then developed was impairing release accuracy

perceptibly. Targets of the size of a medium freighter, seen broadside, gave consistent azimuth control at ranges up to four miles. By inactivating the navigation-correcting elements at moderate range to avoid violent last-minute corrections, the aircraft was consistently able to pass over vessels following straight courses. The improvement in reliable signal quality with respect to the *AN/APG-4* was striking; this was apparently due in part to freedom of the superheterodyne receiver from detector unbalance and in part to increased transmitter power.

Tables VI.-3 and VI.-4 list pertinent electrical and mechanical characteristics of the equipment, including the gyro-stabilizer unit.

TABLE VI.-3

Operating Characteristics of Azimuth–Determining Radar Bombing System

Radar Equipment AN/APG-6(XN)	
Transmitter Frequency:	515 megacycles/sec.
Transmitter Power:	2 watts
Frequency-Modulation Sweep Width:	4 megacycles/sec. max.
Modulation-Sweep Frequency:	Approx. 120 cycles/sec.
Antenna-Switching Frequency:	Approx. 60 cycles/sec.
Operating Altitudes: (Bombing)	40-400 feet
Range:	Approx. 4 miles max.
Azimuth Sensitivity:	±¼ to 1 degree depending on range
Power Source Requirements:	5 amps. at 27 volts d-c
Number of Tubes:	25
Gyro Stabilizer and Servo System MX-247/APG(XN-1):	
Antenna-Servo Tracking Speed:	15 degrees per second
Antenna-Servo Tracking Accuracy:	±3/8 degree
Antenna-Servo Torque:	35 inch-pounds
Navigation Ratios:	2 to 1 or 3 to 1
Target-Tracking Speed:	2 degrees per second
Power Requirements:	5 amperes at 27 volts d-c

TABLE VI.-4

Physical Characteristics of Azimuth-Determining Radar Bombing System

Unit	*Dimensions inches*	*Weight pounds*
Radar Equipment *AN/APG-6(XN)* (Incl. Shock Mount)	18 wide 10½ deep 7 ¾ high	26.3
Transmitting Antennas (two in Yoke)	32 wide 21 deep 20 high	11.0
Receiving Antenna	12 wide 19 deep 20 high	5.1
Altitude Compensation Unit *SA-28/APG*	3¼ wide 6¾ deep 3¼ high	1.8
Gyro Stabilizer *MX-247/APG(XN1)*	8 wide 6¾ deep 11 high	19.8
Antenna Servo	7 wide 7¼ deep 3¾ high	8
Control Box	4 wide 2¾ deep 4 high	1.5
PDI	3¼ dia 4 deep	2.0

5. 1500-MEGACYCLE BOMBING EQUIPMENT *AN/APG-17*

a. *Purpose and Description.* The need to have in reserve an additional operating frequency, as well as a military requirement that antennas be contained entirely within the surfaces of aircraft, led to the development and design for production of the low-altitude bombing radar designated *AN/APG-17* Advent of a requirement for rocket firing prevented its large-scale production, however. While the bombing circuits and tactical operating speeds and altitudes of this equipment are essentially the same as those of the *AN/APG-4,* transmitter mean frequency is increased to 1500 megacycles per second and the cylindrical parabolic antennas shown in Fig. III.-4 are used. These antennas are designed to be contained within the leading edge of the aircraft wing. Installation and maintenance

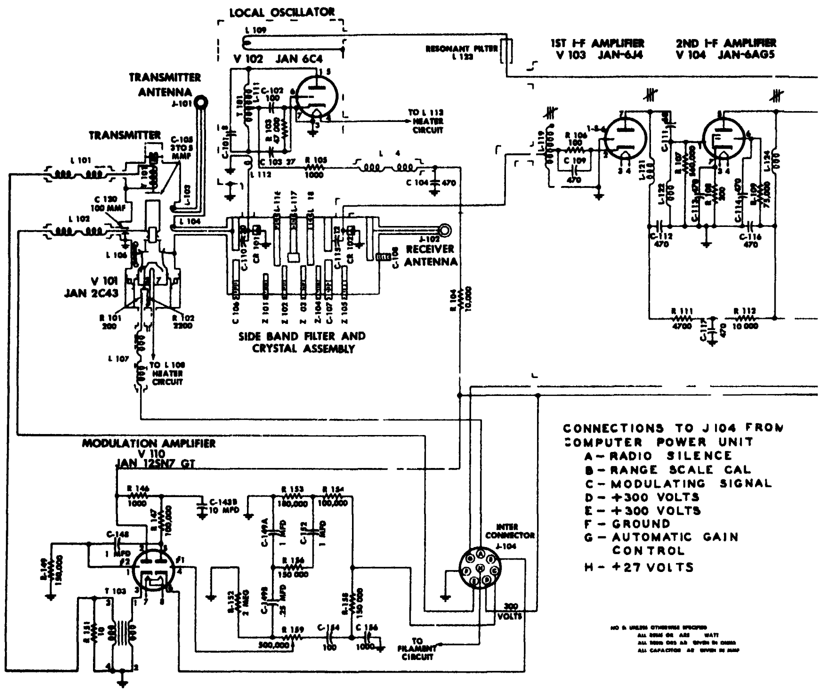
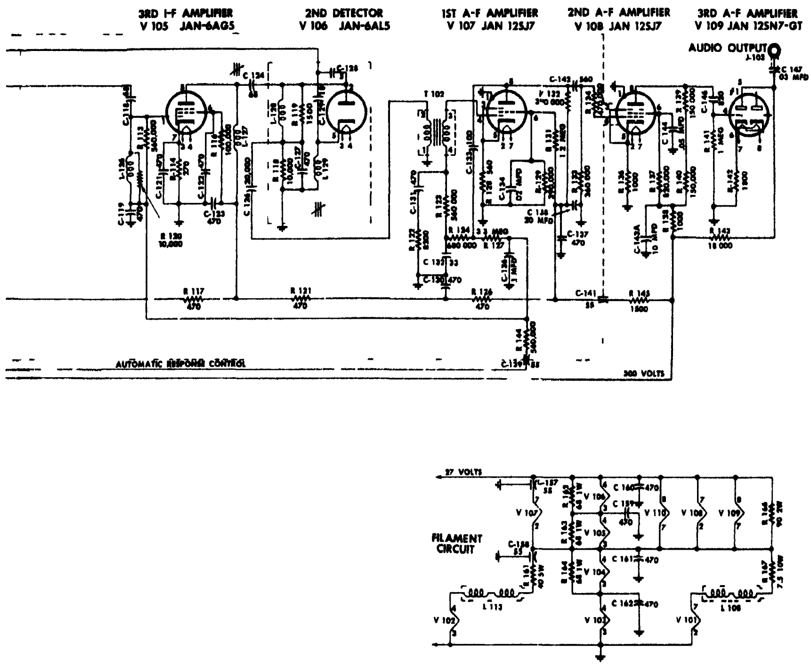


Fig. VI.-21. Circuit diagram of radar

of the equipment in aircraft is facilitated by adoption of a two-unit construction for the production model.

The smaller of the two main units, which is designated *RT-98/APG-17*, comprises the transmitter and receiver up to and including the final audio-amplifier stage. A block diagram of the transmitter and side-band superheterodyne receiver which make up the unit has already been given as Fig. III.-24, and a complete circuit diagram is given in Fig. VI.-21. Compactness, rugged construction, complete shielding, ease of maintenance, and absence of calibrating controls are characteristics of the complete transmitter-receiver unit. Separation of transmitter-receiver functions from computer requirements greatly facilitates adaptation of the equipment to operations other than level-flight bombing. The *AN/APG-17A* and *AN/SPN-2* systems, to be described later in this chapter, are examples of the flexibility of this unit



receiver-transmitter unit *RT-98/APG-17*.

Photographs of the 18-pound *RT-98/APG-17* unit are shown in Fig. VI.-22; a view with front panel removed is included to show the arrangement of sub-assemblies within the unit. One multi-conductor cable, for direct-current heater and plate power as well as modulation-voltage input, and one coaxial cable for low-frequency beat-note signal output, serve to connect this transmitter-receiver unit to the computer-power unit of the equipment. The transmitter, visible at the left front of the interior view of the unit, uses a single *2C43* light-house tube in a grounded-grid resonant-line circuit, as described in section 3a of Chapter III, and shown in detail in Fig. III.-9. The assembly of single-side-band filter, crystal mixer, and crystal first detector, described in detail in section 5b of Chapter III. and shown in Fig. III.-25, is visible at the right front of the unit. Directly behind the side-band filter may be seen a sub-chassis which carries the

120-megacycle intermediate-frequency amplifier and second detector, as well as the audio-input coupling transformer and first audio-frequency amplifier stage. Directly behind the transmitter is the 120-megacycle per second local oscillator, together with inductance-capacitance decoupling filters in the various direct-current leads supplying both transmitter and local oscillator. The sub-chassis at the rear of the interior photograph contains the last two stages of audio-frequency amplification, as well as the triangular and pulse wave-forming and mixing circuits for obtaining linear frequency modulation (see Fig. III.-12), and in addition the modulator tube and transformer to drive the vibrating-capacitor modulator unit. The vibrating modulator itself, which is built into the transmitting oscillator, has been described in section 4b of Chapter III., and a photograph of it appears as Fig. III.-10.

Fig. VI.-22. Transmitter-receiver unit of *AN/APG-17* bombing radar.

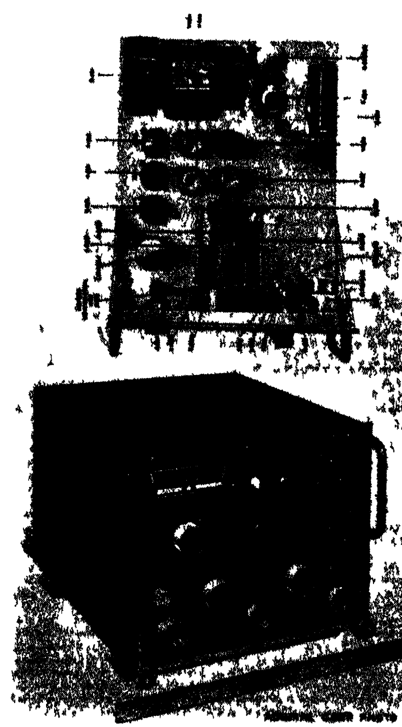


Fig. VI.-23. Computer-power unit of *AN/APG-17* bombing radar.

The 19-pound computer-power unit, which is designated *CP-20/APG-17*, is shown in Fig. VI.-23; a plan view of its chassis is included to show arrangement of parts. Occupying the front central portion of the interior photograph is the mechanism of an *SA-28/APG* automatic altitude compensation unit, which is made an integral part of the *AN/APG-17* rather than an accessory. As described in section 3c of this chapter, the compensation unit includes a follow-up potentiometer driven by a servo motor which is controlled by the limit relays in an **AN/APN-1* altimeter. The servo produces a shaft rotation which is proportional to the altitude of the horizontal center line of the bomb racks above the surface of the terrain. Ganged with the follow-up potentiometer are rheostats used to control modulation sweep width, speed-intercept and residual-range bias, and range-lead bias scale of the bombing equipment, all of which must be varied as functions of altitude. To the right of the altitude-compensation unit are mounted fuses and calibrating controls. The left central portion of the chassis contains tubes for the double limiter, switched counters, release and anti-false-release circuits. Dynamotor, voltage-regulator tube and bomb-release relay are visible at the rear of the chassis.

The counter and bombing circuits of the *AN/APG-17* differ only in detail from those of the *AN/APG-4*, so reference may be made to Fig. VI.-7 for these circuits. Major circuit differences are merely those that result from connecting as integral parts of the *CP-20/APG-17* computer-power unit the accessories that were developed for use separately with the *RT-27/APG-4*. Inclusion of the *SA-28/APG* altitude-compensation unit has already been mentioned. The relays and calibrating controls of the *RE-17/APG*, which permit connection of the *SA-28/APG* to an **AN/APN-1* altimeter without modification of altimeter limit-relay wiring or calibration, are also included in the *CP-20/APG-17* unit, as is the latching booster relay of the *C-141A/APG* release-control accessory. One more such change is inclusion of the relays and circuits of the accessory developed to warn the pilot when release will occur in four seconds. This release-warning accessory is described in section 1b of Chapter VII. and its circuit is shown in Fig. VII.-1.

A further difference between the design of the *AN/APG-17* and that of the *AN/APG-4* is of a mechanical nature. The cam-operated switch from which square-wave modulating voltage is derived is driven in the *AN/APG-17* equipment by a separate speed-regulated motor rather than by the dynamotor. Variations in dynamotor speed in the *AN/APG-4* caused variations in modulation frequency, but the accompaning variation in sweep width resulted in a constant product of sweep width and modulation rate, and therefore in constant radar range sensitivity, as described in section 4c of Chapter III. and illustrated in Fig III.-13. Sweep linearity, however, remains substantially constant only if variations in modulation frequency are slight; this fact determined the use of a separate constant-speed motor for the modulation switch in the *AN/APG-17* equipment. The modulating motor and switch are mounted below the computer chassis and consequently are not visible in Fig. VI.-23.

All operating controls are located in the small unit shown in Fig. VI.-24, designated *C-280/APG-17*, which is connected to the computer-power unit by a single cable.

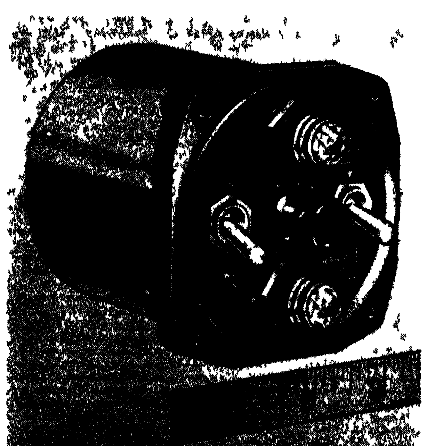


Fig. VI.-24. Control unit of *AN/APG-17* equipment.

This control unit contains the main power switch, a "ready" switch to enable the equipment to release bombs, and the range-lead control. It also contains a lamp to indicate

the armed or ready condition and one to give warning of impending release; both lamps extinguish to indicate release. The same calibrating adjustments as used in the *RT-27/APG-4* and *RE-17/APG* are included in the *CP-20/APG-17* unit.

b. *Operating Characteristics.* Operating procedure in using *AN/APG-17* equipment is identical to that when using *AN/APG-4* with the *SA-28/APG* automatic altitude compensation unit, except that mild evasive action may be maintained until release warning is given. With either equipment accurate automatic bombing is accomplished with the aid of an **AN/APN-1* altimeter or equivalent. Table VI.-5 lists pertinent characteristics of the *AN/APG-17* equipment.

TABLE VI.-5

Characteristics of *AN/APG-17* Bombing Radar

Transmitter Center Frequency:	1500 megacycles/sec.
Power Output at Transmitter:	1 to 2 watts
Maximum Frequency-Modulation Sweep Width:	10 megacycles/sec.
Modulation Frequency:	120 cycles/sec.
Operating Altitudes:	100 to 400 feet
Closing-Speed limits:	100 to 350 knots
Range-Lead Control:	0 to 100 feet
Power-Source Requirements:	27 volts nominal, 5.0 amperes d-c.
Number of Tubes:	18
Total Installed Weight, less Inter-connecting cables:	49.5 lbs.

Transmitter power available in the *AN/APG-17* is approximately ten times that of the *AN/APG-4*, but a corresponding increase in useful sensitivity is not to be expected. Increased losses in radio-frequency transmission lines at the higher frequency waste a large fraction of the increased transmitter output. Since about the same antenna directivity is used in both cases, the effective area of the receiving antenna is much less at the short wave length of the *AN/APG-17* than in the *AN/APG-4* equipment. On the favorable side at the higher frequency is the increased effectiveness as reflectors of flat surfaces and re-entrant

corners on targets. Experimental results are too sparse to permit highly reliable conclusions, but there is at least no definite evidence of superiority of the *AN/APG-17* over the *AN/APG-4* on the score of response to weak targets. This is in contrast to the limited results obtained with *AN/APG-6(XN)*. The freedom of the *AN/APG-17*, like the *AN/APG-6*, from the detector-balance difficulties and instabilities of the *AN/APG-4* does represent a striking improvement in operating reliability. Limited tests[3] have shown the advantage to be gained by using more directive antennas with the *AN/APG-17*, as well as the possibility of operation at higher altitudes.

6. ROCKET-FIRING EQUIPMENT *AN/APG-17A(XN)*

a. *Purpose and Description.* It became evident toward the end of the war that equipment for blind rocket firing would be required. Application of Sniffer principles and methods to meet this requirement was therefore undertaken. Rocket firing was, in fact, considered so important that production of the *AN/APG-17* bombing equipment was withheld to permit its adaptation to use with rockets. Because diving flight at low altitudes under conditions of poor visibility is not ordinarily attempted, automatic rocket firing was only investigated for the case of level flight. Small-caliber service rockets, large-caliber rockets and bombs may all be carried by a single aircraft on a single mission; each requires rather different apparatus adjustments. It was intended that the production rocket-firing equipment *AN/APG-17A* should be made capable of operating at will with any one of the three types of missile, merely by operation of a selector switch. To expedite development, however, the experimental prototype was only made able to choose between two missiles: rockets and bombs. The end of the war found this equipment ready for production design; its development was stopped at that stage.

The rocket-firing system is built into the same two major units as the *AN/APG-17*, a transmitter-receiver and a computer-power unit. The only external modifications to these are, on the computer-power unit, addition of a receptacle for connection to the aircraft rocket-firing circuits and alteration of the labels for the calibrating controls. Modifications within the transmitter-receiver

unit, which are the following, are slight. Some elements in the modulation wave-shaping circuit have altered values to improve sweep linearity. The audio-amplifier input transformer is replaced in the *AN/APG-17A* by a three-section, *m*-derived high-pass filter with a strong rejection point, to give very rapid yet well damped cut off below the pass range. The feed-back network is omitted from the first audio stage and an idle triode is put to use as a fourth audio-amplifier stage. Otherwise, the transmitter-receiver unit is unchanged, so will not be discussed further. Modifications within the computer-power unit are extensive. The appearance of the complete system (less associated altimeter) is shown in Fig. VI.-25. Normal *AN/APG-17* antennas, of the type shown in Fig. III.-4, are used without modification. Only the control unit has been replaced by an entirely new structure, shown in Fig. VI.-26.

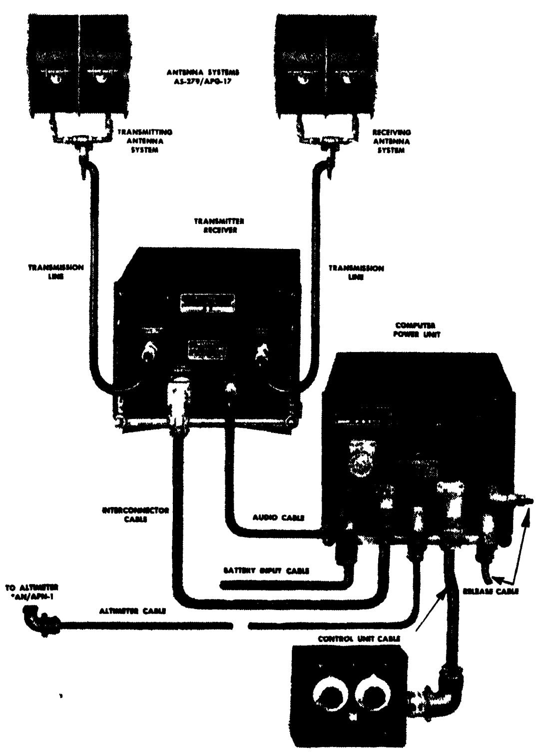


Fig. VI.-25. Complete f-m radar system *AN/APG-17A* for automatic rocket firing.

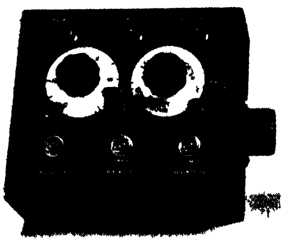


Fig. VI.-26. Control unit of *AN/APG-17A(XN)* rocket-firing radar.

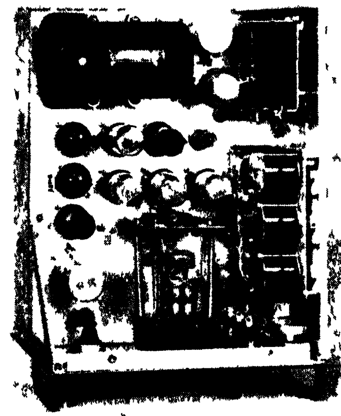


Fig. VI.-27. Interior view of computer-power unit of *AN/APG-17A(XN)* radar.

Internal arrangement of the modified computer-power unit is shown from above in Fig. VI.-27, which may be compared with Fig. VI.-23. The altitude-compensation servo unit again appears at the front center of the photo. To its right are now two banks of three double potentiometers each, used in adjusting the computer to match the characteristics of various aircraft and missiles; their graduated dials are viewed through holes in the side panel to which they are mounted. The tubes of the computer are again at the left and center of the photo, while from left to right at the rear are the plate-supply dynamotor, voltage regulator and power-supply filter, and release and anti-false-release relays. A block diagram of the computer-power and control units is given in Fig. VI.-28 and a functional circuit diagram of them in Fig. VI.-29. As indicated in the block diagram, the apparatus may be broken down into three portions: a set of operating controls and accessories, a release computer operated by the radar signal, and a group of controls capable of adapting the computer to meet various particular conditions of operation. The main computer channel differs from that of the *AN/APG-4* mainly in that it includes a symmetrical speed counter (see section 3a of Chapter IV.) and a separate range counter,

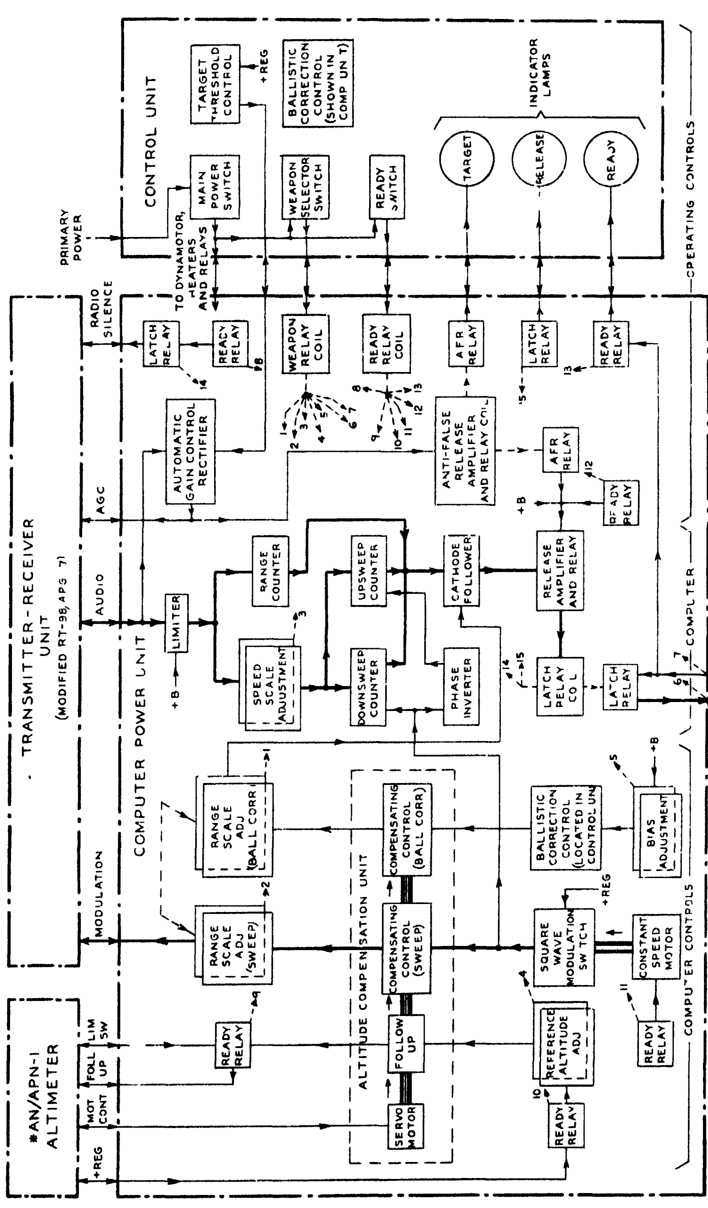


Fig. VI.-28. Block diagram of computer-power and control units of *AN/APG-17A(XN)* equipment.

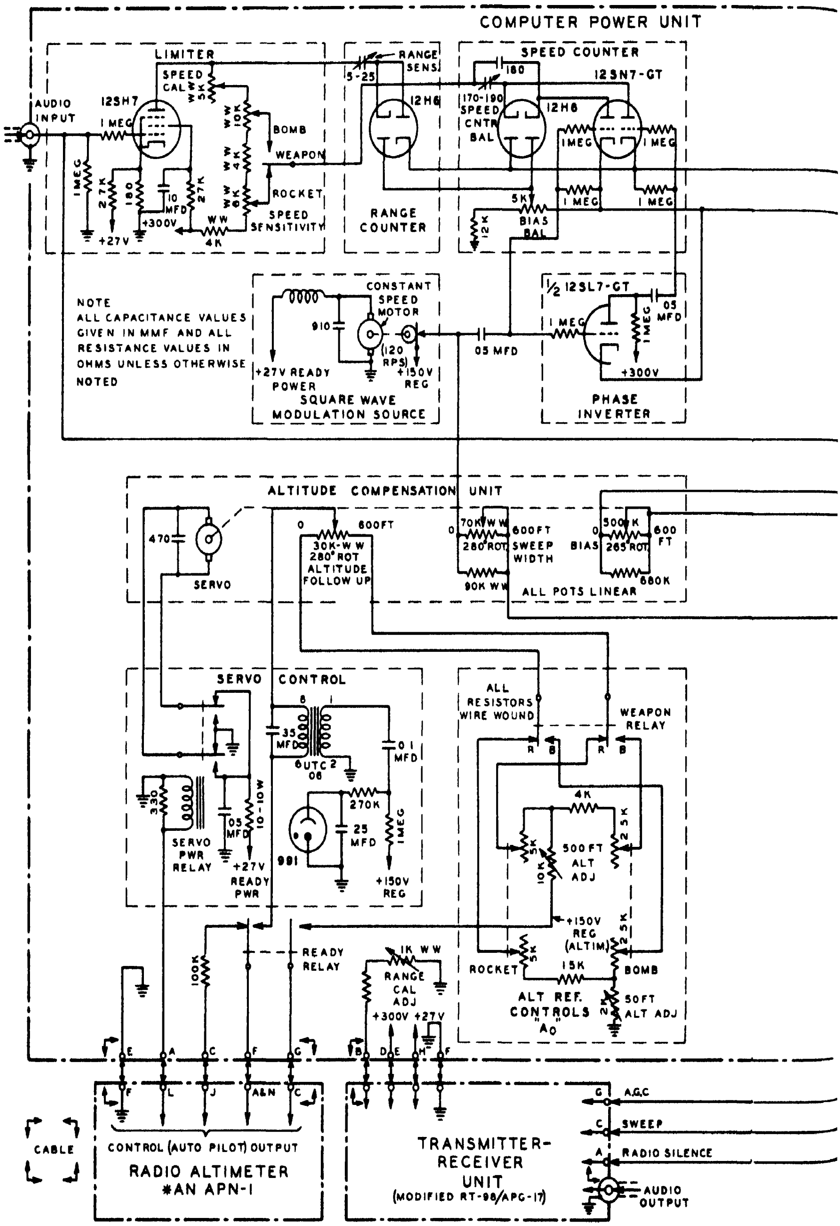
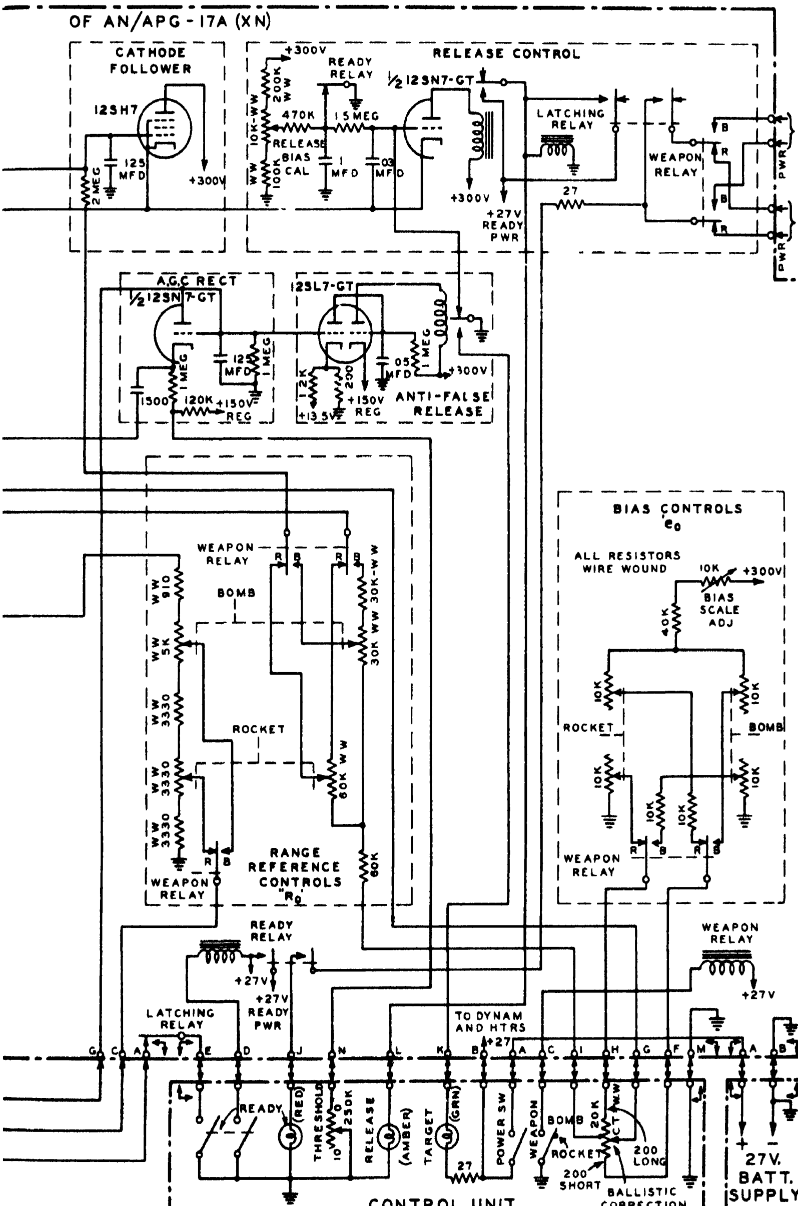


Fig. VI.-29. Functional circuit diagram of computer and



controls of automatic rocket-firing radar *AN/APG-17A(XN)*.

both supplied with square-wave beat-note signal from a common single limiter and both working into a common cathode follower.

Like the *AN/APG-17,* the *AN/APG-17A* includes in the computer-power unit the equivalent of accessory units *SA-28/APG, RE-17/APG* and *C-141A/APG* (see section 3c above), the safety switch and indicator lamps of the last being placed in the control unit. Unlike the *AN/APG-17* but like the *C-141A/APG,* the *AN/APG-17A* provides a lamp indicator that lights upon release and then is kept lighted by a self-latching relay until the system is manually reset. Radio silence is maintained until the safety or ready switch is thrown by hand to the armed position, and is automatically restored by the latching relay immediately a release occurs.

The ready switch in its safe position, beside providing radio silence, de-energizes a relay and thereby removes power from the release-relay circuits and from the altitude-compensation servo motor, as well as stopping the modulation source. The ready relay also controls follow-up connections to the altimeter, so that normal limit-light operation is obtained whenever the *AN/APG-17A* is not armed. A red ready-warning lamp, powered from the aircraft armament circuit selected for use, lights when the ready relay is actuated. With judicious choice of points of connection to the armament circuits, therefore, lighting of this lamp indicates not only that the radar equipment is turned on and made ready to operate, but also that the aircraft circuits for the selected weapon are armed and ready to operate. The ready relay also exerts an anti-false-release action on the release-relay tube, delaying full arming for a sufficient time to prevent false releases caused by switching transients. After each release, the release lamp must be extinguished and the latching relay opened by returning the ready switch manually to its safe position.

No pilot warning was developed for the rocket-firing system, because of the urgent requirement for completed equipment. An anti-false-release device with a two-stage direct-current amplifier actuating a relay is used instead. This lights a target-indicator lamp whenever the received signal exceeds a chosen threshold level, and discriminates

positively against release by disturbances only slightly below that threshold. For most effective operation on small targets, the anti-false-release threshold should be set for the prevailing state of the sea surface; a lockable threshold adjustment is therefore accessibly located in the control unit. Release is only to be expected if both the red warning lamp and the green target indicator are lit. A two-position weapon-selector switch in the control unit completes the operating controls; this actuates selector relays which set up the computer for operation with either rockets or bombs as required.

The really important differences between *AN/APG-17* and *AN/APG-17A* lie in the controls provided for the characteristics of the computer. Four potentiomenter controls and the connections to the aircraft armament circuits are provided in duplicate, as indicated by the double boxes in the block diagram of Fig. VI.-28. These controls are manually preset before flight to adapt the equipment to the characteristics of the missiles in use. One set is used for bombs and the other for rockets; because of the different characteristics of the two missile types, these two sets of four controls are exact duplicates in function only. Selection between missile channels is accomplished by weapon-selector relays; these relays act only to select one of each pair of duplicate elements. Separate control of range and speed sensitivities is a characteristic of the *AN/APG-17A*. Duplicated manual sensitivity controls are provided for the balanced counter, for speed only, which has already been mentioned. Sensitivity of the separate range counter is not varied in operation.

As in earlier equipment, variation of range sensitivity with altitude is required, but in rocket firing the scale of this variation must be varied as well, in accordance with the choice of missile and aircraft used. Duplicate controls for range scale are therefore provided. The basic altitude to which range-sensitivity compensation is referred must be adjusted according to choice of aircraft and missile, and other duplicate controls are provided in the altitude follow-up circuit of the servo for adjusting reference altitude. Improved accuracy of automatic altitude compensation is attained by use of an oscillating

servo of the sort described in section 5c of Chapter IV., instead of the on-off-reverse type of servo used in the *SA-28/APG* and the *AN/APG-17*. A very low-frequency glow-lamp oscillator is used to force small oscillations of the servo motor.

Cathode-follower voltage in the computer should be held constant at release for all operation to provide accurate counter action; relay-tube bias is therefore not varied. Counter-load return bias must on the other hand be quite widely varied, but basic bias need only be adjusted according to choice of aircraft and missile; no altitude compensation of basic bias is needed. Duplicate bias-adjusting controls are provided, making a total of four duplicated controls. Range increments are required to compensate for variations in rocket-propellant temperature and for observed peculiarities in ballistics of particular rocket ammunition, so a lockable manual "ballistic correction" control is provided in the control unit. This produces further bias changes which are added to those from the basic bias-adjusting controls. In order that the range-increment scale of the ballistic-correction control may always read correctly, its bias-increment output is compensated in accordance with changes of range sensitivity of the radar. One control for this bias-increment compensation is ganged with the main radar-altitude-compensation control driven by the altitude servo, while another control is in duplicate and is ganged with the duplicated presettable radar range-scale controls. Each of the four set-up controls appearing in duplicate is provided with a graduated scale for setting to predetermined values.

As in other equipments, calibration is necessary to permit reasonable tolerances on components and operating conditions. Calibration is somewhat more complicated in the case of the *AN/APG-17A* than for other systems, because of the wider range of conditions to be met. Beside the calibration requirements usual in other f-m radar equipment, correct scale readings of the set-up controls must be assured. Voltages applied to the two ends of the reference-altitude adjustment and altitude-servo follow-up circuit are adjustable, to calibrate the altimeter and

servo for correct readings of reference-altitude (A_0) control scales and of the automatic altitude-above-reference (A-A_0) indication of the compensation unit. Gain of the modulation amplifier is adjustable for calibration of the range-scale controls to read correctly. Range-counter capacitor value is adjustable for calibration to a standard range-counter sensitivity. Limiter plate-current swing is adjustable for calibration of the speed-sensitivity control scales. Bias-divider current is adjustable to calibrate the readings of the duplicate bias-control scales, at zero ballistic-correction setting. No calibration for the ballistic-correction scale is provided, nor are means provided for adjusting the shape of the altitude-compensation curve. An adjustment is provided for setting release voltage to a standard value. Provision is also made for balancing the speed-counter capacitors to cancel out speed-counter response to range, as well as for balancing speed-counter diode biases to maintain zero range sensitivity when counter-input voltage swing is varied.

b. *Operating Procedure and Characteristics.* Operating procedure for the *AN/APG-17A* is generally similar to that for the predecessor bombing equipments, but adaptation to a variety of missiles necessitates some differences. At any time before taking off, the four graduated controls of the rocket channel of the computer-power unit must be adjusted to the correct readings, as tabulated for the rocket and aircraft in use, and the bomb-channel controls must be adjusted to settings proper for the class of aircraft in use. Before arrival in the target area, it is desirable if possible to fly into the wind at operating altitude with no target and with armament circuits safe, but with the equipment well warmed up and with radar system armed and in full operation. Under these conditions, the threshold adjustment on the control unit is set so that the target light just remains extinguished in the maximum maneuver to be used in the final approach; the ready switch is then returned to safe position. This will ensure the best operation against small targets that the condition of the sea surface permits. Shortly before attacking, the effective temperature of the rocket propellant should be estimated and the corresponding ballistic

correction set on the control unit. The weapon-selector switch must of course be set for the missile chosen.

For rocket firing, vertical speed or acceleration must be carefully avoided by making as long and accurate a level approach as is feasible, at or near rated military speed of the aircraft. Any desired altitude within the operating range of the equipment for the chosen missile may be used for attack. Upon beginning the approach, with heading controlled by any available aiming data, all circuits for the chosen missile, as well as the *AN/APG-17A*, must be armed. Release will then occur automatically. If rockets are fired, bombs may then be released in the same run by immediately throwing the ready switch to safe position, throwing the weapon-selector switch from rocket to bomb position, and throwing the ready switch again to its armed position. For such dual operation, both rocket and bomb circuits of the aircraft will be armed upon beginning the attack. The ready switch is always to be returned to safe position as soon as possible after a release.

Directly observable operating characteristics will generally be connected with quality of signal. Except for the slowest rockets and largest targets, maximum altitude for reliable operation is determined by working range of the radar. Useful range will only exceptionally exceed 2000 yards, and with a poor target or rough sea may be under 1000 yards. The quality of signal may be judged by behavior of the target-indicator lamp. If this lamp comes on well in advance of firing and remains on steadily until after firing, accurate results may be expected. If it comes on late and flickers on and off at the time of firing, the target is inadequate and must be attacked at lower altitude.

Test firing of three different types of rocket has shown that the methods developed for adapting the equipment to operation with a variety of missiles have been successful. A moderately large number of radar firings made with one particular type of rocket has indicated that the overall dispersion in automatic firing with this equipment is not significantly greater than the natural ballistic dispersion of the rockets themselves. Signal quality during the firing tests indicated that decidedly greater transmitter power than the one to two watts available would have been

desirable, as would greater antenna directivity.

c. *Adaptation to Ballistic Requirements.* Internal characteristics of the *AN/APG-17A(XN)* are controlled by the requirement, developed in section 5 of Chapter V., that at firing

$$^{RF}/R_o - {}^{S}/S_r = 1 - {}^{S_m}/S_r - {}^{R_bF}/R_o , \tag{VI.5}$$

where F is a generally useful altitude-compensation function given by

$$F = \frac{125(A - A_o + 865)}{865(A - A_o + 125)} . \tag{VI.6}$$

S, R and A are slant closing speed, slant range, and level-flight altitude at the instant of release (or, more exactly, at an instant prior to release by the total time lag of the system). Mid-operating speed S_m is a constant which is characteristic of the firing aircraft and speed increment S_r is a constant characteristic of the rocket, while both reference altitude A_o and mid-speed release range at reference altitude, R_o, are characteristic of the combination of aircraft and missile. The radar system is arranged to close its release relay when

$$k_R h_R R - k_S h_S S = e_1 - e_o , \tag{VI.2}$$

with k_R, k_S, h_R, and h_S representing as usual range and speed sensitivities of radar and counters respectively; e_o is cathode-follower grid voltage for zero range and speed and e_1 is follower-grid voltage for relay operation. Values of these sensitivities and voltages must be chosen to make (VI.2) directly represent (VI.5).

In equation (VI.5), the total manually adjustable range increment R_b for ballistic correction includes both a portion R_T dependent upon departure of propellant temperature from 60 degrees Fahrenheit, and a residual portion $-R_r$ to compensate for range equivalent of radio-frequency lines and for displacement of antenna location behind missile-rack location on the aircraft. It may also include an arbitrary correction R_1 to meet special tactical requirements. The sign of the total ballistic correction R_b is so chosen that a positive or "long" correction would move a point of impact initially on the target to a point

beyond the target. This convention, by which an impact beyond the target would be considered a positive range error, is the opposite of the "range lead" convention of earlier f-m radar equipments but is in agreement with the more usual convention of missile range errors.

System capabilities might have been harmonized with operating requirements, as in earlier systems, by using constant counter sensitivities in all cases and adjusting sweep width and bias to meet all requirements; this would have required a moderate range of sweep widths and only a single switched counter, sensitive to both range and speed. Examination of (VI.5) will show, however, that range beat-note frequency for release at mid speed S_m (with zero total ballistic correction) would in that case vary widely with missile characteristic S_r. This would greatly widen the required audio-amplifier pass band and make discrimination against feed-through signal and short-range sea-return signal much more difficult.

In the *AN/APG-17A*, the requirements have been met instead by the alternative method of using separate range and speed counters and adjusting the speed-counter sensitivity in accordance with the value of S_r, while retaining a constant range-counter sensitivity. Under these conditions, (VI.5) indicates that release at mid speed ($S = S_m$) and with zero total ballistic correction R_b will always occur for the same range frequency, independently of aircraft, rocket and altitude. This permits a decidedly narrower amplifier pass band, sufficient only to allow for the moderate departures from mid speed to be expected in operation, for necessary ballistic corrections, and for an adequate duration of signal to permit transients to die out before firing. The price is that a decidedly wider but still acceptable variation in sweep width is required to meet all conditions.

Again, numerical values will serve to illustrate magnitudes of various quantities involved in design and operation. Design factors, almost all arbitrarily set at given values for reasons of convenience, are assembled in Table VI.-6. Insertion of these design values into equation (VI.2) will show immediately the term-by-term agreement with (VI.5) so obtained. Operating conditions,

dictated by properties of rockets and aircraft as well as by equipment-design factors, are given for a typical case in Table VI.-7. The large net counter output at the instant of firing rockets is quite characteristic and is in marked contrast to the small net output when releasing bombs. An accurate counter-load resistor is therefore necessary for rocket firing.

TABLE VI.-6

Design Characteristics of *AN/APG-17A(XN)* Rocket-Firing Radar

Radar:	
Radio Frequency:	1500 megacycles per second
Radar Speed Sensitivity k_s:	3.05 cycles per second per foot per second (5.15 cycles per second per knot)
Modulation Frequency:	120 cycles per second
Maximum Permissible Modulation Sweep:	20 megacycles per second
Useful Range of Modulation Sweep Width:	1½ to 14 megacycles per second
Sweep Width:	$12,290F/R_0$ megacycles per second (R_0 in feet)
Radar Range Sensitivity k_R:	$6.00F/R_0$ kilocycles per second per foot.
Range Beat Frequency at Release at Mid Speed:	6.00 kilocycles per second
Audio Gain Characteristic:	Increases 10½ decibels per octave from 2.6 to 6.8 kilocycles per second, decreases 6 decibels per octave above 8.5 kilocycles, falls off at 105 decibels per octave from 2.4 to 1.8 kilocycles.
Computer:	
Cathode-Follower Grid Voltage e_1 for Release:	100.0 per cent, or reference voltage (1/3 of plate-supply voltage)
Bias Voltage e_0 to Counter Load:	$28.0+72S_m/S_r+72R_bF/R_0$ per cent of reference voltage.
Range Counter Sensitivity h_R:	12.0 per cent of reference voltage (4.0 per cent of plate-supply voltage) per kilocycle per second

TABLE VI.-6 (Cont'd)

Overall Range Sensitivity $k_R h_R$:	$72.0F/R$ per cent of reference voltage per foot
Range Counter Output for Release at Mid Speed:	72.0 per cent of reference voltage
Speed Counter Sensitivity h_S:	$14.0/S_r$ per cent of reference voltage per cycle per second
Overall Speed Sensitivity $k_S h_S$:	$72.0/S_r$ per cent of reference voltage per knot.

TABLE VI.-7

Typical Operating Conditions for *AN/APG-17A(XN)*

Medium-Speed Rocket	
Slow Aircraft	
Aircraft Mid Speed S_m:	215 knots
Operating Air-Speed Range:	215±15 knots
Wind and Target Speed Range:	± 40 knots
Rocket Characteristic Speed S_r:	750 knots
Reference Altitude A_o:	70 feet
Reference Range R_o:	2085 feet
Propellant Temperature Correction R_T:	-8 feet per degree Fahrenheit above 60° F.
Flight Altitude A:	250 feet
Altitude above Reference A-A_o:	180 feet
Altitude-Compensation Factor F	0.496
Mid-Speed Firing Range R_o/F:	4220 feet
Modulation Sweep Width:	2.91 megacycles per second
Radar Range Sensitivity:	1.42 cycles per second per foot
Range-Counter Sensitivity:	0.012 per cent per cycle per second
Overall Range Sensitivity:	0.0171 per cent of reference voltage per foot
Radar Speed Sensitivity:	5.15 cycles per second per knot
Speed-Counter Sensitivity:	0.0186 per cent per cycle per second

TABLE VI.-7 (Cont'd)

Overall Speed Sensitivity:	0.096 per cent of reference voltage per knot.
Bias Voltage to Counter Load (Zero Ballistic Correction):	48.6 per cent
Ballistic-Correction Bias Increment:	0.017 per cent per foot
Range Frequency at Point of Firing:	6000 cycles per second
Range-Counter Output at Firing (at Mid Speed, zero Ballistic Correction):	72.0 per cent
Speed Frequency at Mid Speed:	1107 cycles per second
Speed-Counter Output at Mid Speed:	20.6 per cent
Net Counter Output at Mid-Speed Firing:	51.4 per cent
Total Counter Output at Firing (Including Bias):	100.0 per cent of reference voltage (1/3 of plate-supply voltage).
Minimum Flight Altitude:	100 feet
Maximum Useful Range:	6000 feet (approx.)
Maximum Altitude (Range Limitation):	470 feet (approx).

Factors determining the audio-frequency band to be amplified for rocket firing are indicated in Table VI.-8. The limiting case of the slowest rocket and aircraft and the lowest altitude to be used is the basis of this table; other rocket-firing conditions require less audio band. Bombing, however, requires extension of the pass band to include ideally 2.1 to 13.5 kilocycles. A characteristic of the type suggested, as roughly approximated in the equipment built, is found to eliminate almost completely interference of any reasonable level caused by low-frequency microphonics, altitude signal, short-range sea return, or low-frequency feed-through signal. Whether this gain-frequency characteristic results in extra counts from fixed error, as does the simple differentiating characteristic (see Chapter IV., section 2g), has not been certainly determined. Because the lowest frequency passed to the counter in the *AN/APG-17A* is many times the radar modulation frequency, such extra counts can not cause very serious

error even if present. Allowance can easily be made for them if necessary in setting the bias voltage applied to the counter load. No net extra count results from counter switching in the case of the symmetrical speed counter used.

TABLE VI.-8

Factors Determining Audio Characteristic for Rocket Firing

	Slow Rocket Slow Aircraft
Minimum Altitude:	100 feet
Reference Altitude A_o:	72 feet
Rocket Characteristic Speed S_r:	550 knots
Total Closing-Speed Range:	215 ± 55 knots
Reference Range R_o:	1715 feet
Altitude-Compensation Factor F:	0.844
Mid-Speed Firing Range:	2030 feet
Total Ballistic-Correction Range:	± 720 feet (± 240 yards).
Firing-Range Increment from Speed Variations:	± 205 feet
Limits of Firing Range:	1105 to 2955 feet
Range Frequency at Mid-Speed Firing (Zero Ballistic Correction):	6000 cycles per second
Radar Range Sensitivity:	2.96 cycles per second per foot
Limits of Range Frequency at Firing:	3265 to 8735 cycles per second
Limits of Speed Frequency:	825 to 1390 cycles per second
Audio Frequency Band Needed at Firing:	2440 to 10125 cycles per second
Duration of Good Signal Needed before Firing:	2 seconds
Maximum 2-second Range Increment:	455 feet
Maximum Range-plus Speed Frequency:	11470 cycles per second
Desired Audio Characteristic for Rocket Firing:	Rises 10 decibels per octave from 2.4 to 11.5 kilocycles per second, falls off below 2.4 and above 11.5 kilocycles as rapidly as possible without "ringing".

The *AN/APG-17A*, which requires altitude compensation of sweep width according to the simple analytical expression (VI.6), provides a good example of a useful method for shaping the compensation characteristic. Fig. VI.-30 shows the circuit used, r being a linear rheostat that has its rotation, and therefore resistance, made directly proportional to altitude above reference A-A_0, which may here be called A_1. Values of resistance r_θ and altitude above reference $A_{1\theta}$ at a chosen standard angle θ of electrical rotation, together with the electrical angle θ_0 of its mechanical zero stop, completely define this rheostat.

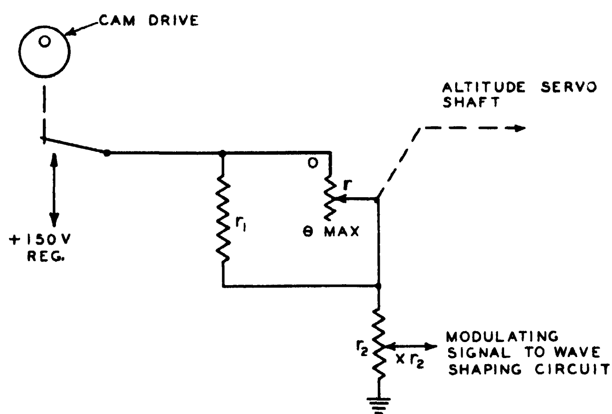


Fig. VI.-30. Modulation-control circuit for altitude compensation of *AN/APG-17A(XN)*.

The rheostat is shunted by a fixed resistor of value r_1 and this combination is in series with another fixed resistor r_2. Output to a very high-impedance load is taken from a variable tap at a fraction x of r_2, which thus serves also as an output-voltage divider.

The total square-wave output-voltage swing E is

$$E = ir_2 x = E_o \frac{r_1 r_2 + r_2 r}{r_1 r_2 + (r_1 + r_2) r} x. \tag{VI.7}$$

In terms of altitude above reference, this is

$$E/E_o = x \cdot \frac{r_2}{r_1 + r_2} \cdot \frac{A_1 + A_{1\theta} r_1/r_\theta}{A_1 + \frac{r_2}{r_1 + r_2} \cdot A_{1\theta} r_1/r_\theta} \tag{VI.8}$$

Except for the output-divider ratio x, this expression shows

the circuit to have exactly the properties required by equation (VI.6) for the altitude-compensation function F. By setting the divider ratio x, in terms of reference range R_o, to a value $R_{o_{min}}/R_o$, the complete circuit may be made to give directly the desired quantity F/R_o; this divider is the manual range-scale adjustment mentioned earlier.

To meet specific compensation requirements, it is only necessary to provide proper resistance ratios. This permits a reasonable tolerance on the resistance value of rheostat r, which is a primary element in determining the accuracy of the entire system, provided fixed resistances r_1 and r_2 are made to have values properly related to the actual characteristics of r. The altitude servo of the *AN/APG-17A(XN)* operates over altitudes above reference A-A_o (or A_1) from zero to 600 feet, and is so adjusted as to produce just 220 degrees of rotation at an altitude 500 feet above reference. Compensation should therefore have the proper form, illustrated in Fig. VI.-31, if r_1/r_{220} is made $\frac{865}{500}$ and r_2/r_{220} is made $\frac{865}{500}\Big/\left(\frac{865}{125}-1\right)$.

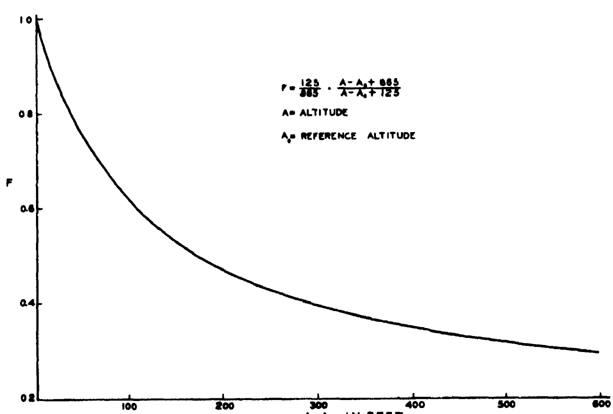


Fig. VI.-31. General-purpose altitude-compensation function.

Actually, the ratios required for proper curve shape are found empirically to be somewhat different from these (1.64 and 0.292 respectively), probably as a result of finite impedances of voltage source and load. Earlier equipments had the modulation-generating switch at the grounded end of

the compensation circuit, and still greater disturbances of the compensation-curve form were then produced by a monitoring bleeder resistor shunted across the switch (see Fig. VI.-7). This was not finally harmful but did complicate design conditions. Resistance r_2 in the altitude-compensation circuit of the *AN/APG-17A(XN)* equipment includes potentiometers which provide ranges for R_o from 680 to 1020 feet for bombing and from 1530 feet to 3060 feet for rocket firing.

Scale reading of the ballistic-correction adjustment must be maintained correct, as adjustments of r and of tapping point on r_2 vary radar range sensitivity. This is done by interposing, between the center-tapped potentiometer which provides incremental bias for ballistic correction and the counter-load return lead to which bias voltage is supplied, a compensating circuit exactly similar to the one (shown in Fig. VI.-30) interposed between the modulation source and the modulation wave-shaping circuits. Corresponding control elements in the modulation-compensating and bias-compensating circuits are ganged together mechanically, as indicated by dotted-line mechanical connections in Fig. VI.-29. R_o values in the range provided for rocket firing happen to be just three times those provided in the bombing range (as far as the bombing range extends). Duplication of the r_2 potentiometer in the bias-compensating circuit (with the potentiometer range in the bomb channel suitably adjusted) therefore permits a single graduated scale to be used on the ballistic-correction potentiometer. This scale indicates directly either range increments between plus and minus 240 yards for rockets or plus and minus 240 feet for bombs.

7. SPEED MEASURING EQUIPMENT *AN/SPN-2(XN-1)* FOR CARRIER CONTROL OF APPROACH (*CCA*)

a. *Purpose and Requirements.* Aircraft landing on carrier vessels must hold their air speed within narrow limits during approach. Too low an approach speed may cause premature stalling, while too high speed may result in overshooting the arresters on the flight deck. Serious speed error in either direction is likely to result in destruction of the aircraft.

Under conditions of good visibility, the Landing Signal Officer (*LSO*) of the carrier vessel judges approach speeds of aircraft by observation of their flight attitude. The *LSO* controls approach and landing, by hand signals to the aircraft pilot, in accordance with this and other data which he obtains by direct visual observation of the aircraft.

Under conditions of poor visibility, control by the carrier of the landing approach of aircraft requires an integrated system of radio aids; actual landing is still controlled by the *LSO* after he has the aircraft in view. In one experimental system for Carrier Control of Approach (*CCA*), altitude information is supplied directly to the pilot by the limit lights of his **AN/APN-1* altimeter, range and azimuth are supplied to the controller aboard the carrier by a special precision pulse radar [*AN/SPN-3(XB)*], and approach-speed data is supplied to the controller by the special *AN/SPN-2(XN-1)* equipment described in this section. The approach is directed by the controller, who issues orders to the aircraft pilot by radio telephone.

For f-m radar measurement of range only as in the **AN/APN-1* altimeter, or of both range and speed with range predominant as in the *AN/APG-17A* rocket-firing equipment, it is clearly desirable that range-beat frequency shall exceed speed-beat frequency. Speed can be determined separately under this condition by a symmetrical switched counter, but elimination from the speed data of the large range signal then requires critical counter balance. For determination of speed only, independently of range, it is clearly desirable that speed-beat frequency shall exceed range-beat frequency under all expected conditions of operation. Speed may then be reliably determined by a simple unswitched counter.

In the *AN/SPN-2* equipment for speed measurement, speed-beat frequency is kept above range-beat frequency by reducing frequency modulation, and thereby range-beat frequency, substantially to zero. This equipment is therefore not intentionally an f-m radar at all. It was developed, however, by merely reducing the modulation sweep width of standard f-m radar apparatus as far as practicable, so may properly be discussed here.

Radar equipment *AN/SPN-2* measures only the speed of the aircraft relative to the carrier. This differs from the air speed of the aircraft by the air speed of the carrier, which is the speed of the wind over the carrier. Correction of the radar output for wind over carrier is therefore necessary to provide direct indication of air speeds of aircraft. Operation is required, on carrier-based aircraft at ranges up to one nautical mile, for air speeds of 40 to 140 knots and for winds over carrier of zero to 40 knots. Beside quantitative indication of air speed, a qualitative indication of whether that speed is greater or less than an adjustable pre-assigned value must be provided.

The fact that a large portion of the echoing area of an approaching aircraft is a spinning propeller leads to considerable difficulty in obtaining accurate, stable speed indications. Special means are therefore provided in the *AN/SPN-2(XN-1)* equipment to minimize this difficulty. Means are also provided to prevent misleading indications in the absence of an adequate radar signal.

b. *General Description.* The *AN/SPN-2(XN-1)* equipment uses two weather-proofed, horizontally polarized single-dipole antennas with small reflecting "hats", each mounted in a 4-foot diameter parabolic main reflector; such antenna units are sharply directive. Three major units which, in addition to the antennas, compose the speed-measuring system are: a transmitter-receiver unit, a computer-power unit and an indicator unit. The antennas are mounted, pointing aft and well separated, at the stern of the carrier vessel and just below the flight deck. The transmitter-receiver unit is mounted in a waterproof housing directly between the antennas, to minimize length of r-f transmission lines and consequent losses.

Transmitter-receiver and computer-power units do not differ significantly in appearance from the *RT-98/APG-17* and *CP-20/APG-17* units of the *AN/APG-17* bombing equipment, shown in Figs. VI.-22 and VI.-23, from which they were modified. The transmitter-receiver unit is modified only by removal of the vibrating-capacitor modulator and its driving circuits, and by alteration of the frequency characteristic of the beat-note amplifier. A rigid diaphragm bearing a tuning trimmer capacitor replaces the

vibrating modulator in the transmitter. Circuit changes in the computer-power unit are extensive. An entirely new indicator unit, shown in front and top-rear views in Fig. VI.-32, carries all operating controls. This unit is designed for mounting in the *AN/SPN-3* radar console, directly available to the approach controller.

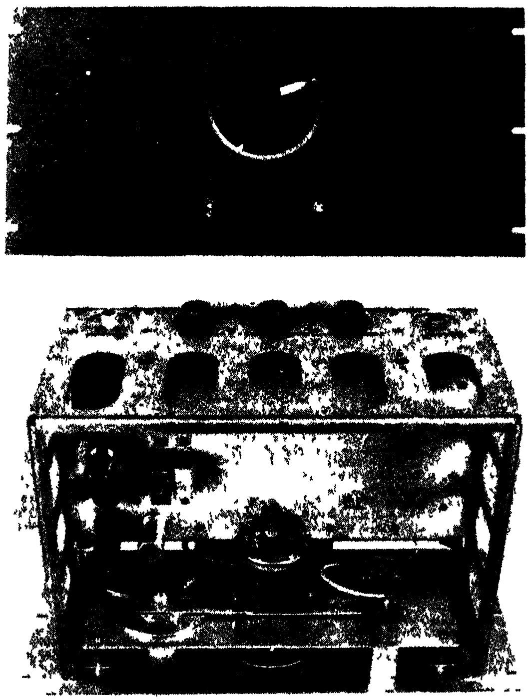


Fig. VI.-32. Indicator unit of speed-measuring equipment *AN/SPN-2* (*XN-1*).

The side-band superheterodyne receiver of the *AN/APG-17* is used intact. Reduction of the side-band filter pass band to the minimum required by oscillator drift, by removal of transmitter modulation, makes filter tuning much less critical. The beat-note amplifier characteristic is made flat from 200 to 700 cycles per second and falls off rapidly at lower and higher frequencies. High gain at these low frequencies makes elimination of microphonics important and so requires use of improved tube types for i-f amplifiers and second detector. Reduction of frequency modulation of the transmitter and amplitude modulation of the local oscillator to the lowest attainable levels is of major importance.

A block diagram of the computer-power and indicator units alone is given in Fig. VI.-33; Fig. VI.-34 is a functional circuit diagram of the same two units. The main signal channel of the computer-power unit, shown in heavy lines at the top left of the latter figure, is composed of a double limiter and a sharply selective tunable filter. A counter and servo system for tuning the filter is also shown as included in the computer-power unit; this will be discussed later. Beside these elements and the usual dynamotor and glow-regulator power supply, the computer-power unit contains two audio-signal rectifiers, for automatic gain control of the intermediate-frequency amplifier and for actuation of a "hold-off" or indicator-disabling relay when signal strength falls below a chosen threshold. Separate rectifiers with different time constants are used, to permit rapid action of the hold-off relay during short fades without affecting the normal slow action of the automatic gain control.

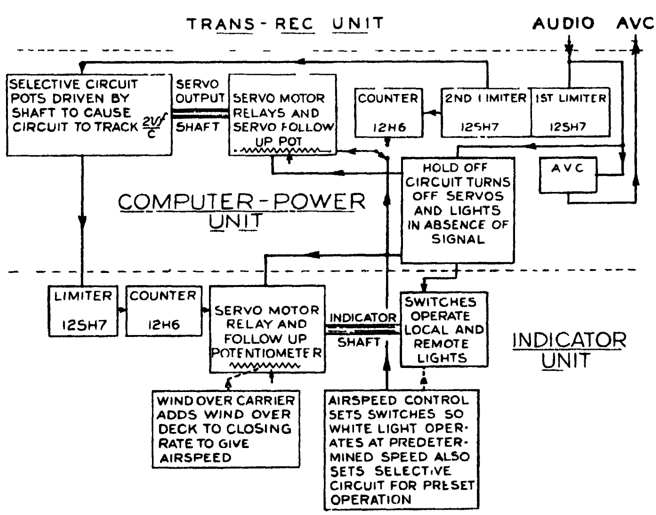


Fig. VI.-33. Block diagram of speed-measuring radar *AN/SPN-2 (XN-1)*.

In the indicator unit, the main-channel signal from the selector actuates a single limiter and a simple servo-balanced null counter like that of Fig. IV.-20. Action of the indicator servo is made smooth by operating it in a rapidly vibrating condition. Servo vibration is excited

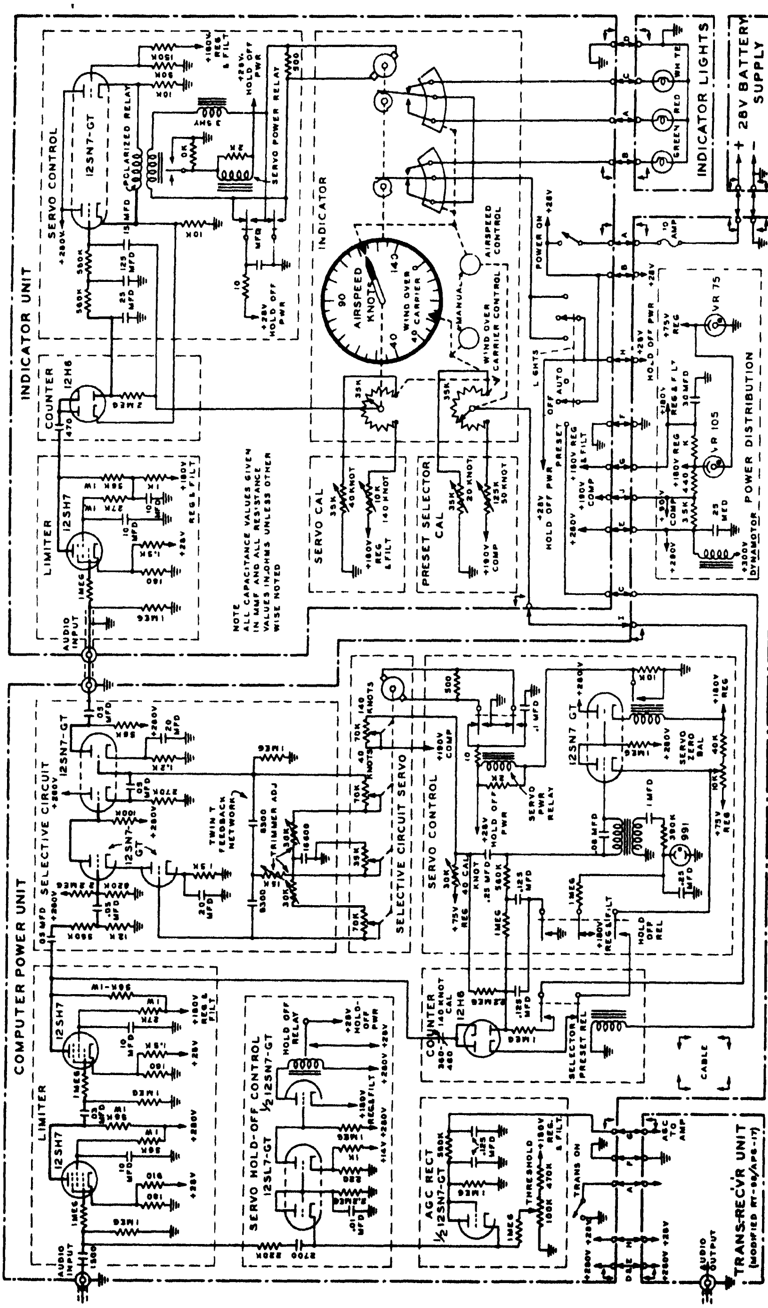


Fig. VI.-34. Functional circuit diagram of computing and indicating portions of *AN/SPN-2(XN)*.

by applying the voltage across the servo-motor armature to a balanced polarized relay that is used to control the direction of motor rotation in accordance with the sense of counter unbalance. This feed-back connection is made through a series inductor to an auxiliary winding on the polarized relay, with such polarity that the system of motor and relay is self-oscillatory at a frequency (near 16 cycles per second) controllable by the value chosen for the series inductor.

Since the null counter and the follow-up potentiometer are linear, the angular position of the indicator-servo shaft with respect to the potentiometer winding is directly proportional to the relative speed of the target with respect to the radar. Total angular rotation of the servo with respect to the fixed speed-indicating dial is made proportional to air speed of the aircraft target by manually setting the frame and winding of the potentiometer to an angular position that is proportional in turn to the air speed of the radar, which is the speed of the wind over the carrier. The servo-driven pointer therefore directly indicates target air speed against the main scale of the indicator dial. A movable index shows, against an auxiliary wind-over-carrier scale, the angular setting of the winding of the follow-up potentiometer. The operator makes this setting with a "Wind Over Carrier" knob. Friction braking of the wind-setting shaft prevents operation of the servo from disturbing its settings. Because a wind setting may need to be made with the servo stationary at one of its limits of travel, and the required setting may not be compatible with that servo position, spring relief has to be provided in the coupling from wind-over-carrier knob to follow-up potentiometer frame to prevent mechanical damage.

An additional potentiometer on the indicator-servo shaft, and a shaft extension and mounting space for a synchro-transmitter unit as well, are provided to permit remote air-speed indication. A remote indicator at the *CCA* controller's station actuated by the wind-measuring system of the carrier vessel is desirable to insure correct adjustment of the wind-over-carrier control.

The indicator-servo shaft rotates two cams which actuate

switches mounted on a carriage in turn rotatable manually about the servo shaft. An index reading against the main air-speed scale of the indicator marks the position to which these switches are set. When the indicator air-speed reading is within ±2½ knots of the switch setting, made by the operator with the "Air Speed" knob of the indicator unit, a white "Speed Correct" lamp is lighted by the switches. At indicated speeds below the preset 5-knot range, a red "Speed Too Low" lamp is lighted, and at speeds above that range a green "Speed Too High" lamp is lighted. Sets of limit-indicator lamps may be located wherever desired, for example at the station of the *LSO*, but none are provided within the indicator unit. The width of the correct-speed light range is easily set by mechanical adjustment of the cams on their shaft. Indicator lamps are automatically extinguished and the indicator servo is stopped by the hold-off relay unless an adequate signal is received.

c. *Reduction of Propeller Modulation.* Modulation of the reflection characteristics of the aircraft by motion of its propeller distorts violently the wave form of the Doppler-frequency beat between transmitted and received radar signals. Such distortion represents complex modulation of both amplitude and frequency of the beat-note signal. The modulation frequency is dependent upon speed of propeller rotation and number of propeller blades, but is usually near 80 cycles per second for aircraft approaching a carrier landing. Limiter action reduces though it does not necessarily eliminate the amplitude modulation, which may be extreme, but it does not affect the undesired modulation of beat-signal frequency.

Slow action of the indicator servo averages out variations of the Doppler-beat frequency and so reduces unsteadiness of indicated speed which results from propeller modulation. It is found, however, that the average frequency indicated tends to be too high in the presence of propeller modulation. This is thought to be caused by dissymmetry of the audio-amplifier gain characteristic with respect to the unmodulated Doppler frequency, which is usually in the neighborhood of 300 cycles per second. Such dissymmetry weights the upper modulation side band more

heavily than the lower side band and so may cause high readings. The required amplifier characteristic is fixed by the need to suppress low-frequency response to microphonics, sea return, and random noise, while passing all desired speed-beat frequencies.

A sharply selective filter tuned to the unmodulated frequency of the desired beat signal will not only reduce all modulation side bands but will also render them symmetrical. This should facilitate the proper averaging of propeller-modulated speed data, but poses the problem of maintaining the filter tuned to the incoming signal. The filter used in the computer-power unit of the *AN/SPN-2(XN-1)* system to suppress propeller modulation employs an amplifier with selective degenerative feed back. This type of filter, using a twin-T feed-back network,[4] is used because of the ready availability of its components and its ease of tuning, even at very low audio frequencies.

Tuning of the filter is accomplished by a servo motor, which drives a follow-up rheostat and three other rheostats in the three resistive arms of the twin-T network. The filter resonates at a frequency which is proportional to $1/\sqrt{r_p r_s C_p C_s}$, where r_p, r_s, C_p, and C_s are, respectively, the total resistance and capacitance values in the parallel and series arms of the twin-T network. These resistances are each composed of a fixed resistor in series with a linear rheostat on the servo shaft, the ratio of fixed to total resistance being the same for all resistive arms of the network.

The follow-up circuit is also composed of a fixed resistor in series with a linear rheostat on the servo shaft, again with the same ratio of fixed to total resistance as in the feed-back network. With this follow-up circuit connected across a constant-voltage supply, the current flowing in the circuit and therefore the voltage across the fixed resistor is proportional to $1/r$, where r is the total resistance in circuit. That is, for any setting of the tuning-servo shaft, the follow-up voltage produced is directly proportional to the frequency to which the filter is tuned. Both follow-up voltage and filter frequency vary in hyperbolic fashion with servo-shaft rotation, but both do so in just the same way. This useful result requires

equality of the ratios of fixed to total resistance in all four circuits adjusted by servo rotation.

Arrangement of the servo to run so as to balance its own follow-up voltage output against a control voltage permits automatic tuning of the filter. The tuning-control voltage may be supplied either by counter-output current flowing in a load resistor and proportional to beat-note frequency, or by a manually adjustable source in the indicator unit. Since no negative-voltage supply is available in the system, manual control is made possible by arranging the servo to balance with the grid of its controlling amplifier at a fixed positive voltage rather than at ground potential. The source of follow-up voltage is especially compensated against voltage-supply variations. Vibration of the tuning servo to insure smooth operation is forced by a glow-lamp oscillator.

Variation of speed of an aircraft in landing approach is slow. Filter tuning by the counter-controlled servo may therefore be slow also. Slow filter response to changes in received-signal frequency prevents the filter from locking to and following any momentarily strong side frequency, such as may be produced in an occasional interval of especially intense propeller modulation. The tuning servo is stopped by the hold-off relay during periods of inadequate signal, and the counter-output smoothing capacitors are disconnected at the same time to preserve their stored voltage until signal returns and so to prevent needless transient operation of the servo after fades.

Manual control of filter tuning is obtained from a potentiometer adjusted by the "Air Speed" setting knob of the indicator unit. The frame and winding of this potentiometer are turned by the "Wind Over Carrier" setting knob. Control voltage fed to the tuning servo, and consequently the filter frequency, is therefore proportional to speed of the aircraft relative to the radar when flying at just the preset air speed.

The operating condition is selected by a three-position switch on the indicator unit. In the "Off" position, limit lights are extinguished but the indicator servo operates normally and the tuning servo automatically tunes the filter to the incoming signal. In the "Auto" position, the

filter is still tuned automatically to the incoming signal but the limit lights as well as the indicator dial are operative. In the "Preset" position, the limit lights operate and the filter is fixedly tuned for the speed at which the "Speed Correct" lamp has been preset to light.

d. *Operating Characteristics.* Except for setting in the value of wind over the carrier necessary to convert measured relative speed to air speed, the equipment is fully automatic in operation. Once turned on, the equipment indicates target air speed directly whenever a signal sufficient for good operation is received. In the "Auto" or "Preset" condition, absence of limit-light operation denotes lack of an adequate target; speed indications then observed are to be disregarded. Some characteristics of the *AN/SPN-2(XN-1)* system are given in Table VI.-9.

TABLE VI.-9

Characteristics of *AN/SPN-2(XN-1)* Speed-Measuring System

Radio Frequency:	1500 megacycles per second
Transmitter Power:	1.5 watts
Power Input:	5 amperes direct current at 26 volts (nominal)
Total Beam Width of Antenna between Half-Amplitude Points:	22° in azimuth, each antenna; 16° in elevation, each antenna
Speed Sensitivity:	5.15 cycles per second per knot
Air Speed:	40 to 140 knots
Wind Over Carrier:	0 to 40 knots
Maximum Range on Carrier-Based Aircraft:	1 nautical mile
Amplifier Characteristic:	flat (± 1 db.) 200 to 600 cycles per second; falls 30 db. per octave below 200 cps.; falls 11 db. per octave above 1000 cps.
Filter Characteristic:	60 cycles/sec. wide at 3 db. down

Tests over land and water have indicated a maximum range of two and a half miles on carrier-based types of aircraft, as a result of the large directive power gain of the antennas used. This ensures sufficient signal at

the one-mile maximum range for which the equipment was designed. Overall accuracy tests with a shore installation, while not conclusive, have indicated accuracy to be within one knot in the region of 90 knots.

Discrepancies of the order of 4 to 15 per cent have been found between *AN/SPN-2* speed measurements and the aircraft's true air speed as derived from indications of dynamic air-pressure increment. Where such discrepancies appeared, the aircraft's air-speed indicator was checked by timed flight over a measured course and was found invariably to be in error. In all cases the error was in the direction to cause the aircraft to be travelling at a higher speed than indicated. No significant errors in *AN/SPN-2* indications were found.

Ship-board tests made with preliminary equipment indicated the need of means for minimizing propeller-modulation difficulties, as described above and incorporated in the final equipment. Ship-board tests of the final equipment have shown these measures to be effective, the equipment giving steady and accurate speed indications despite propeller motion. The ship-board tests also showed the advantage gained because aircraft are only taken aboard with the carrier headed into the wind. This advantage comes from the fact that sea return from the windward side of waves is much less than that from their lee side; the radar pointed astern sees only the windward side of the waves when aircraft are landing. Since the speed of the waves relative to the radar can produce Doppler frequency shifts up to perhaps 150 cycles, strong sea return might cause considerable difficulty. It is therefore important that amplifier gain should fall off as rapidly as possible below the minimum desired Doppler frequency from aircraft.

It has been apparent from all tests that elimination of frequency modulation makes radar equipment much less critical as to its surroundings. In fact, the unmodulated equipment can be operated successfully when installed in locations that would be quite impossible for frequency-modulated radar because of strong range-beat signals from nearby objects.

Continuous-wave Doppler-shift speed measurement is of particular interest because of the extreme sensitivity and

accuracy theoretically attainable. Radio frequency can be determined and maintained to practically any absolute accuracy desired. Velocity of propagation is known very accurately and is constant to a high degree. Beat frequency can be measured very accurately if desired. The only serious limitation seems to be in the definiteness of the speed to be measured itself. Rapid measurements to high accuracy require the use of the highest practicable radio frequencies, in order to obtain high beat frequencies.

8. NOTATION AND REFERENCES

a. *Notation*. The algebraic notation listed alphabetically below has been used in this chapter.

A	Altitude of aircraft above terrain.
A_0	Reference value of altitude.
A_1	Altitude above reference value.
$A_{1\theta}$	Altitude above reference for shaft angle θ.
C	Circuit capacitance (usually with identifying subscript).
e	Voltage output of circuit.
e_0	Bias voltage applied to counter load.
e_1	Total counter-output voltage at which a relay operates.
e_R	Comparator-circuit voltage using rightward-pointing antenna.
e_L	Comparator-circuit voltage using leftward-pointing antenna.
E	Modulating-voltage amplitude.
E_0	Supply voltage.
f_a	Frequency of antenna switching.
F	General altitude-compensation function.
h_R, h_S	Counter sensitivities to range and speed beat frequencies.
i	Circuit current.
k_R, k_S	Range and speed sensitivities of radar.
r	Circuit resistance (usually with identifying subscript).
r_θ	Rheostat resistance at shaft position θ.
R	Range or distance between radar and target.
R_0	Reference value of range.
R_1	Range lead or increment for special purposes.

R_b	Total ballistic-correction increment to range.
R_r	Range equivalent of r-f transmission lines.
R_T	Range increment for departure of rocket-propellant temperature from reference value.
S	Relative speed of approach of radar and target.
S_o	Intercept of linear range-speed relation on speed axis.
S_m	Mid-range operating speed of aircraft.
S_r	Speed characteristic of rocket.
Sw	Switch.
T'	Slope of linear range-speed relation at release.
V	Vacuum tube (with identifying subscript).
W	Width of frequency band swept in modulation.
x	Fractional position of voltage-divider tap.
θ	Angular position (electrical) of rheostat arm.
θ_o	Electrical rheostat position at mechanical stop.

b. *References.*

1. D. G. C. Luck: "Sniffer Operation at Increased Altitudes," Report no. 4 under Contract NXsa-35042 (May 24, 1944).

2. D. Blitz: "Automatic Altitude Compensation Unit *SA-28* for Use with *AN/APG-4* (Sniffer), and Adaptation of *AYD* and *AYF* Altimeters for an Altitude Servo System," Report no. 2 under Contract NXsa-35042 (April 22, 1944).

3. V. D. Landon and O. M. Woodward: "Investigation of Various Antennas and the Feasibility of Operation up to 800 feet Altitude on the *AN/APG-17*," Report no. 22 under Contract NXsa-35042 (April 17, 1945).

4. R. M. Walker and H. Fleisher: "The Use of a Twin-T Network in a Selective Frequency Amplifier, with Special Applications," Radiation Laboratory Report no. 737 (May 19, 1945).

CHAPTER VII.

ACCESSORY CIRCUITS AND METHODS OF CALIBRATION

1. RELEASE-WARNING ACCESSORY

a. *General.* Automatic bomb-release equipment, of the Sniffer type described in the preceding chapter, makes the operations required of the pilot in low-altitude level-flight bombing very simple. It does not aid him in judging the stage to which his approach has progressed, however. An accessory to provide warning of impending bomb release[1] is therefore very useful, since it enables the pilot to continue moderate evasive action to the latest possible stage of his approach and to plan an expeditious withdrawal.

Two types of warning circuit have been developed. One of these gives a warning signal at a definite time interval before release, while the other warns at a definite distance before release. Either type requires only the addition of two relays and no tubes to the normal equipment. Both operate by changing the release calibration of the system, so that the normal release action of the Sniffer takes place twice in succession, once under calibration for actual release. Only the second Sniffer release is permitted to reach the bombing circuits of the aircraft.

Either type of circuit, whether time or distance, provides a warning interval that is compensated for altitude, for closing speed, and if desired for range-lead setting. The aircraft must of course be flown within normal Sniffer altitude and speed limits, both at warning and at release. No vertical velocity is permitted the aircraft at release, but this restriction is not important for the warning operation. Any increase in altitude occurring between warning and release will shorten the warning interval, while a decrease in altitude will lengthen it with either circuit. Warning of course need not be a precision operation.

b. ***Time Warning.*** The basic range-speed-altitude relation by which a relay can be actuated at a definite time interval before an aircraft, in level flight and bearing an f-m radar, passes over its surface target (see section 3a of Chapter V.) is easily set up for any time interval desired. An interval greater by any definite amount than the time of fall of a bomb from the altitude of flight can be produced by suitable choice of frequency-modulation sweep width and of bias applied to the release-relay amplifier.

Upon actuation of the Sniffer release relay at the time chosen, during approach to a target, a warning lamp may be lighted and a second relay may be actuated and latched in. This second relay alters the sweep to the increased value required for later release of a bomb to strike the target, changes the release-tube bias to the normal value for bombing, and connects still a third relay to be actuated and latched by reopening of the Sniffer relay. Increased sweep then causes the Sniffer counter output to rise and the release relay to reopen, actuating the third relay. The third relay thereupon reconnects the release relay so as to operate the bomb-release mechanism, through a self-latching booster relay if desired, when next actuated. When the target range finally decreases to the proper bombing value for the altitude and speed used, normal bomb release occurs. All relays must be reset manually after release.

A pilot-warning accessory for the standard *AN/APG-4* and *SA-28/APG* combination may be built as a separate, self-contained small unit, replacing the usual *C-141A/APG* control unit. The circuit of such a unit, as arranged to provide a four-second warning, is shown in Fig. VII.-1 in proper relation to the *SA-28/APG* and the main Sniffer unit. Sweep width is suitably altered by changing both the fixed shunt on the servo-driven altitude-compensating linear rheostat in the modulating circuit and the fixed series resistor across a part of which modulation output from that circuit appears. The change in bias to allow for warning calibration of speed intercept and residual range is approximated sufficiently well by a fixed voltage change. Compensation for range lead is simple but requires extra

connections to the circuits of the *SA-28/APG*, so is not included in Fig. VII.-1. The warning circuit used in the *AN/APG-17*, with its internal *SA-28/APG* unit, includes compensation of warning for range lead.

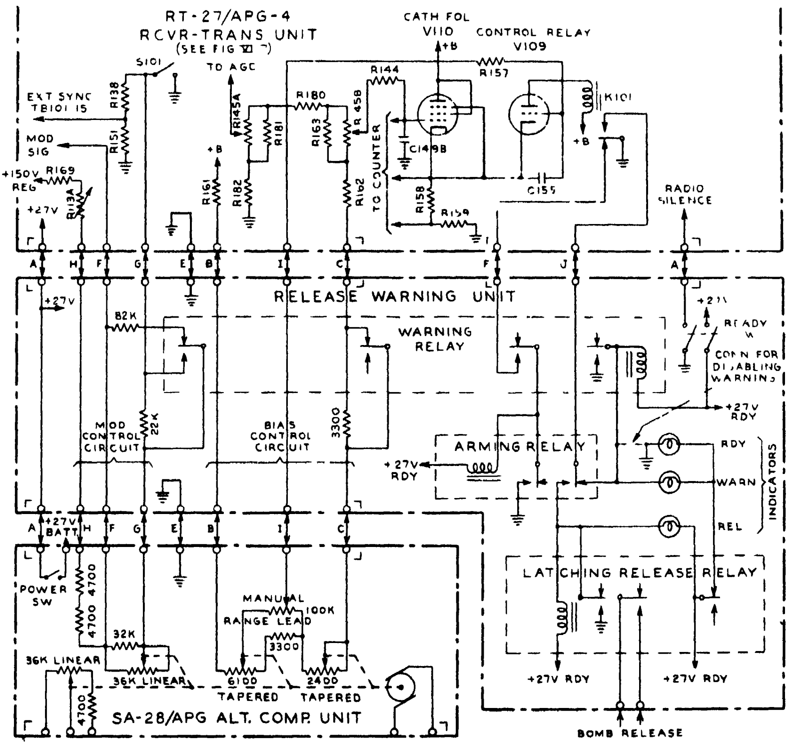


Fig. VII.-1. Pilot's four-second release-warning accessory unit for *AN/APG-4*.

The ready or arming switch and the latching relay of the *C-141A/APG* are retained in the warning unit shown, as are the ready and release indicator lights. Only the warning and arming relays and a warning light, as well as three fixed resistors, are added. If the connection shown dotted in the figure is made, it will pre-actuate the warning circuits and so permit normal Sniffer bomb-release operation without previous warning. A warning and release-control unit will require connection to the altitude-compensation unit and to the corresponding circuits of the Sniffer, as well as to the release circuits of the Sniffer and to the bombing circuits of the aircraft. The complete warning

circuit of Fig. VII.-1 is included within the normal computer unit of the *AN/APG-17* equipment, where it operates as here described except for its additional feature of range-lead compensation.

c. ***Range Warning.*** As has been noted in connection with residual range and range lead, altering the bias voltage applied to the relay-tube grid of an *AN/APG-4* by an amount proportional to the range sensitivity of the system will change range at release by a constant increment, independently of speed and altitude. This makes possible a warning signal operating at a range greater than release range by a fixed amount. Operation on this basis involves beginning the approach with a bias value suitable for the desired warning distance from the release point, switching automatically to the bombing bias and to the bombing circuits as warning is given at that distance, and finally releasing the bomb normally at the proper range.

A voltage proportional to radar range sensitivity is available in the form of the modulation-frequency square-wave output of the altitude-compensation unit. To use this voltage as a release-bias increment for warning, two things must be done: filtering must be removed from the relay-tube input so that the square-wave peaks can operate the release relay, and counter output must be reduced so that the bias thus obtained will represent a sufficient range increment. The capacitor used for relay-tube input filtering in normal bombing provides a convenient means for coupling in the square-wave bias increment used to shift calibration to the warning condition; the square-wave voltage is thus simply added to the normal steady bias. Reduction of the plate voltage supplied to the limiter and to the bias-divider chain provides the needed reduction in counter sensitivity. Protection against false release due to transients immediately after warning is obtainable by applying intentionally a strong positive transient of adequate duration to the relay amplifier, in the process of changing from warning calibration to bombing calibration.

Fig. VII.-2 shows a circuit for warning at a range 1200 feet greater than that for release. Because the relay-tube filter capacitor is not accessible externally,

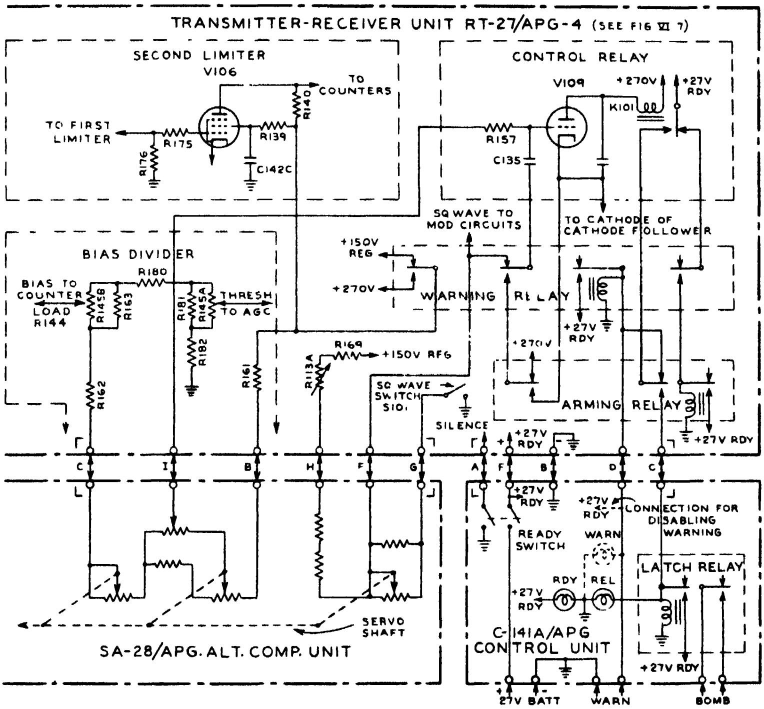


Fig. VII.-2. Pilot's 1200-foot release-warning modification of *AN/APG-4*.

a self-contained separate accessory unit for range warning is not feasible, and the circuits of the figure must be incorporated into the regular Sniffer computer. The latching booster relay for bomb release may be in the external *C-141A/APG* unit, as shown.

d. *Comparison of Warning Methods.* Warning circuits of both fixed-time-interval and fixed-range-interval types have been fully developed and tested in trial bombing flights. Both operate as expected. Both reduce the vulnerable straight and level portion of the approach to the distance travelled between warning and release. It is possible to interpret either type of warning as a time interval; if the range interval is fixed by the circuit, the equivalent time interval will depend upon the speed of the aircraft. It is alternatively possible to interpret either type of warning as a range interval; if the time interval is fixed by the circuit, the equivalent range

interval will depend upon the speed of the aircraft. In either case, no serious confusion is likely to result, since a pilot engaged in military operations will habitually fly a single type of aircraft having only a narrow range of operating speeds. There is therefore little to choose between the two methods as regards convenience in use.

The warning sequence must always operate before release can occur. Its use may therefore affect the reliability of the equipment, particularly in the case of small targets. Release can only occur on targets large enough to have previously actuated the warning. In the fixed-range type of warning, radar range sensitivity does not change and audio frequencies at warning are much greater than at release. The sweep-width reduction made in giving the fixed-time type of warning, on the other hand, is just right to oppose the increase of range and produce an audio frequency at warning approximately equal to that present at release.

Range warning must take place at frequencies which are usually above the response peak of the audio amplifier, so a stronger target than is necessary for accurate release is usually required for correct warning. On a target which is of approximately the marginal size for normal release, warning will not occur until range has fallen to the value giving an audio signal at the peak-response frequency of the amplifier. Release will then take place normally.but the warning interval will have been too short. Still smaller targets will not operate the warning, but would not give accurate Sniffer releases without the warning circuits either.

Time warning requires consideration of two cases. If the marginal target is fixed by spurious radar reflections, such as altitude signal and sea return, the decreased sweep for warning will reduce the frequencies of these spurious signals. The sloping amplifier characteristic will therefore reduce the spurious-signal level, and indeed by approximately the same amount as increased target distance reduces the desired-signal level. Size of the marginal target will therefore not be affected in this case by use of the warning circuit. If the marginal target for

accurate release is fixed by thermal or other noise not affected by sweep width, the fact that target range at warning may be as great as 2½ times the range at release will require a target up to 25 decibels more effective to give fully accurate warning. The target must be in this case at least 10 decibels more effective when the warning circuit is used, or warning and subsequent release will fail entirely to occur.

2. ANTI-FADING ACCESSORY

a. *Conditions to be Met.* Brief fading of the radar signal at or near the time of release, when bombing with Sniffer equipment, can cause serious error in bomb impact. An accessory to offset the ill effects of fading[2,3,4] should therefore be capable of improving overall bombing accuracy. At the beginning of a normal approach, voltage at the grid of the cathode follower which linearizes the counters of the *AN/APG-4* has a saturated high value, corresponding to counter loading by onset of grid current in the cathode-follower tube. During the approach, this voltage decreases steadily with decreasing range, at a rate which depends upon altitude and speed of approach. When the follower-grid voltage reaches the threshold set by the voltage applied to the relay-tube grid, release occurs. Fading of signal disturbs the steady decrease of cathode voltage and so introduces errors in release. A "memory" device which preserves the steady voltage variation during fades will prevent such release errors.

Two different sorts of fading error can occur. One sort results from strong thermal noise or high-frequency noise due to loose-contact modulation of the feed-through field around the aircraft. Lack of signal in a "noisy" system will result in high follower-grid voltage, caused by counting of the high-frequency noise components, which are emphasized by the high-peaked audio-amplifier characteristic. This effect of random noise was in fact relied upon during early Sniffer tests to prevent release if the equipment were armed without the presence of a target. Fading in this case results in a rapid rise of follower-grid voltage during the fade, followed by decrease to normal voltage after the signal returns. A fade at or just before release with a noisy system delays the decrease to

threshold voltage and so causes a late release and bomb impact beyond the target. A rapid succession of fades may prevent release altogether.

The other sort of fading error occurs in "quiet" systems having especially low random noise. Residual signal in the absence of a target is then of low frequency and is due to short-range sea return, altitude signal or, in the *AN/APG-4*, imperfect detector balance. Follower-grid voltage in the absence of target signal will in such systems be at a minimum, set by counter biases, which is below the threshold for release. During a fade, the follower-grid voltage will fall below the value it would have with signal present, increasing to a normal value quickly after the return of signal following the fade. A fade at or near release with a quiet system will therefore cause the cathode follower to reach release-threshold voltage too soon and the bomb to fall short of the target.

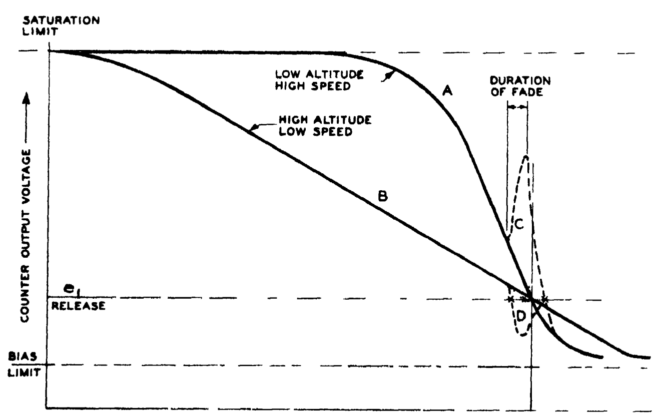


Fig. VII.-3. Effect of fading on counter output.

Fig. VII.-3 shows the variation of follower-grid voltage with time during approach, the full-line curves *A* and *B* representing normal operation with strong, steady signals. Rapid decrease of voltage is characteristic of approaches at high speed or low altitude, while slow decrease is characteristic of low speed or high altitude. Release occurs when the threshold voltage is crossed. Dashed curve *C* represents the effect of a brief fade just before release, during a fast approach made with a noisy system.

A late release during the return to normal voltage after the fade is marked by the cross on dashed curve *C*. Dashed curve *D* represents the effect of a similar fade during a slow approach with a quiet system. The resulting premature release is marked by the cross on dashed curve *D*.

Use of a noise-integrating double limiter (Chapter IV., section 2f) tends to make systems behave in quiet fashion, thus emphasizing low-frequency disturbances, preventing the voltage rise of dashed curve *C* on fading, and favoring false or premature release. The anti-false-release circuit of the *AN/APG-4* disables the negative counter during fades and so prevents the voltage fall of dashed curve *D*; indeed, it substitutes a rise like that of *C*. Premature releases are avoided by this anti-false-release action, but severe fading may cause late release and poor accuracy.

Accuracy may be restored artifically despite fading if two conditions are met. First, the counters must be positively prevented from transferring charge to their output-filtering capacitor (C_0 in the figures of Chapter IV., *C149B* in the *AN/APG-4* circuit of Fig. VI.-7) unless a good signal is available. Second, the charge already on this capacitor must be allowed to leak off during fades as a "memory" discharge at the proper rate to duplicate the fall of output voltage found in normal operation. Duplication during each fade of the actual rate of voltage fall present at its beginning is possible but rather complicated; it is much easier to provide a fixed rate of uniform memory discharge that is representative of normal operating conditions. Dotted curves *C* and *D* of Fig. VII.-3 show voltage variation in fast and slow approaches during a brief fade, with memory discharge at a single fixed rate of average value. The great reduction in disturbance of normal discharge which results is clearly evident by comparison with the dashed curves. Since counting is suspended, the dotted curves do not depend at all upon the noisy or quiet condition of the radar system.

Over the normal altitude and speed range of *AN/APG-4* equipment, actual rate of voltage fall in approach varies from 6 to 43 volts per second. It is therefore difficult to pick a single typical rate, but the short duration of fades at high speeds suggests the greater importance of

the lower rates. Since any fixed memory-discharge rate can only be a crude approximation over any significant range of operating conditions, it is evident that the follower-grid voltage can become seriously incorrect during a long fade. The fixed-rate discharge therefore provides only a short-time memory, and must be prevented from causing markedly inaccurate release during a prolonged loss of signal.

b. *Functions of Circuits.* The set of operations required of an anti-fading memory accessory now becomes evident. When signal is definitely absent, the accessory must definitely maintain a disarmed condition to prevent bomb release. When a usable signal first appears, the accessory must immediately arm the system and place it in condition for a normal approach. Whenever signal fades even momentarily to less than usable level, the accessory must disable both counters, initiate memory discharge of the counter-output capacitor, and initiate timing of the fade. If usable signal returns soon enough, the accessory must terminate memory discharge, reinstate normal counter operation, and reset to zero the fade timer. If signal has not returned by the time that maximum tolerable error may have accumulated, the fade timer must disarm the system and prepare it to start operation afresh when good signal is again received. Equipment capable of performing these functions is easily arranged to provide also a definite indication of the presence of adequate signal.

Fig. VII.-4 shows a circuit developed for use as an anti-fading accessory. Elements normally present in the *AN/APG-4* bear the same reference numbers as are shown in Fig. VI.-7 for that equipment. Elements external to the normal *AN/APG-4* represent the controls and memory of the anti-fading circuit proper; they include as well the functions of automatic gain control and prevention of false release, so replace the elements normally used for these purposes, and make them available for use in the accessory. Three triode tubes and two relays are evidently required in addition to the normal Sniffer elements. Any single relay with a sufficient number of contacts to perform all required operations exhibits a difference between the current at which it closes and the current at which it opens that is

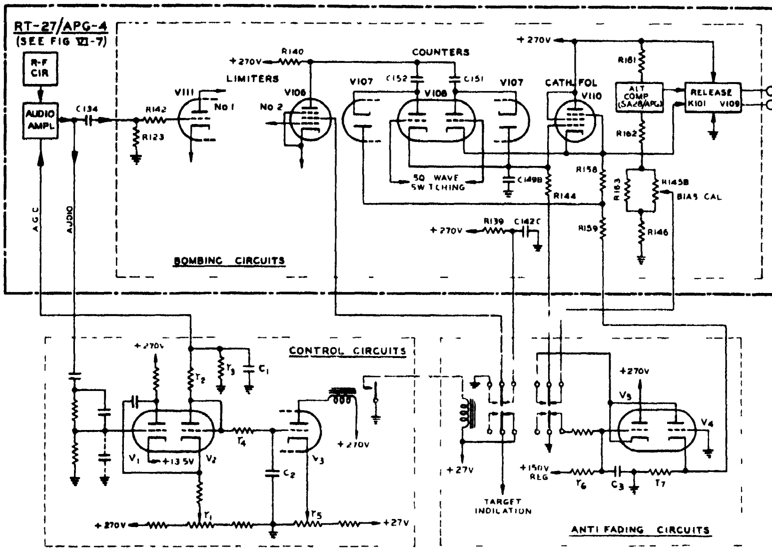


Fig. VII.-4. Functional diagram of anti-fading accessory circuits with short-time memory.

intolerably large for this use. Two relays are therefore used in cascade, the first having simple contacts and small hysteresis.

Audio-amplifier output is fed, through a suitable capacitance-compensated voltage divider to avoid overload, to triode V_1 which is biased to cut off and operates as a class-B amplifier with large output. V_2 is a biased half-wave rectifier operated by the output of V_1; it provides a large, negative direct-current control voltage with magnitude determined by level of radar-beat signal above a threshold set by potentiometer r_1. A portion of this control signal is filtered by r_2 and C_1 and is used for automatic amplifier-gain control, actuated by excess of audio signal over the threshold set. The complete control signal is filtered by r_4 and C_2 and applied to relay-actuating tube V_3; control-signal level for relay actuation is set by potentiometer r_5. By proper choice of the relative values of control-voltage components across r_2 and r_3, and of their associated time constants, relay-operating threshold in fades can be made practically independent of the signal level preceding a fade-out or fade-in. This independence of a-g-c action is important for satisfactory operation.

When adequate operating signal is present, the relays are open as shown and Sniffer operation is entirely unaffected by the presence of the anti-fading equipment. The disarming or anti-false-release tube grid is then held at substantially zero voltage, and a target-indicating lamp is lighted. When the signal fades below the chosen threshold level, the relays close immediately. The counters are then disabled by grounding the screen of the limiter tube, the target indicator is extinguished, and a constant-current memory discharge of *C149B* through triode V_4 is initiated. Voltage on the grid of fade-timing and disarming tube V_5 begins to rise at a rate determined by r_6 and C_3. If signal returns soon enough, the relays open immediately and fully normal Sniffer operation is restored, the counters correcting very rapidly any small error in voltage on *C149B* and cathode-follower grid caused by inappropriateness of the fixed memory-discharge rate. If the fade continues long enough, the voltage on the grid of V_5 will exceed that on *C149B*. V_5 will then act as a cathode follower; its cathode current will overcome the memory discharge and raise the voltage on *C149B*, thus disarming the system and thereafter maintaining it disarmed so long as the relays remain closed.

The plate of grounded-grid discharge tube V_4 is at the grid voltage of the cathode follower and its cathode is tapped onto the cathode-follower load resistor. Plate-cathode and grid-cathode voltages on V_4 therefore have opposite signs and proportional magnitudes. If the grid/plate voltage ratio on this tube is approximately equal to its amplification factor, plate current becomes substantially independent of applied voltage. The value of the constant plate current used for memory discharge can be adjusted within limits by changing the grid/plate voltage ratio on V_4, determined by the tapping point on the follower load set by the value of r_7. Memory-discharge current is of the order of two microamperes.

c. *Operation.* Evaluation of the operation of a device to reduce bombing errors caused by fading is difficult, since it is only operative in border-line cases and a considerable degree of randomness then exists. With good signals, the device does not act; with really poor signals,

accurate releases cannot be obtained in any case. When release is made with memory in a border-line case, it is not possible to determine what would have happened on that particular release without memory action. Real information can only be obtained by statistical analysis of very many border-line releases, some made with and some without memory action.

Because of these difficulties, the limited number of test bombs dropped by an *AN/APG-4* system with an anti-fading attachment has been insufficient to permit a definite evaluation of the usefulness of memory action. Most of the test drops were made using a cruder device[2,3] than the final anti-fading accessory described above.[4] This did not have a constant-current memory discharge but only a resistive leak, nor was its relay-action threshold compensated against disturbance by operation of the automatic gain control; the disarming action was also produced in a less satisfactory way.

Even under these circumstances, it seemed evident in the tests made that bombing accuracy was improved by memory action in a number of instances. This was true especially at higher altitudes, where fading was often rather severe in the region of release range. There was also some indication that conditions occasionally arise in which a fading sequence interacts with the operating sequence of the anti-fading device to decrease bombing accuracy. Relay-action threshold must be set to require a signal good enough for accurate release, if memory is to operate to best advantage; weak targets which might sometimes give a fortuitously accurate release by the Sniffer alone are therefore barred from arming the system at all when the anti-fading accessory is used. Several of the conditions that the anti-fading device was developed to alleviate are adequately handled by the combination of double limiter and anti-false-release circuit, which was added to the standard Snifier while work on the anti-fading accessory was in progress.

Very positive indications were found in the flight tests that a decidedly sharp and carefully set threshold of relay action is essential to make good use of an anti-fading device. This is so because the margin between a threshold

level at which too many targets fail to arm the system and a lower level at which excessive inaccuracy is produced by various disturbances is surprisingly narrow. The tests of the cruder device also indicated the importance of making the improvements incorporated in the relatively untried version here described.[4] This improved form of the accessory is incorporated in the *AN/APG-6(XN)* azimuth-controlling bombing equipment. Values for its circuit constants may be found in Fig. VI.-14, the schematic circuit of that equipment.

3. RANGE CALIBRATION

a. *Variable-Frequency Beat Method.* It has been mentioned repeatedly in discussing various f-m radar systems that empirical calibration is necessary to reduce errors resulting from accumulation of normal tolerances on circuit-component values and operating conditions. This applies particularly to radar range sensitivity k_R as determined by the modulation used. The fact that the average over the modulation cycle of the range-beat frequency depends only on the modulation frequency and the total width of the radio-frequency band swept in modulation, and not on the details of the modulation wave form, has already been pointed out in section 4i of Chapter II. This fact suggests one basic method of range calibration, in which the limits reached by the transmitted radio frequency are measured and the resulting range sensitivity of the radar is calculated.

The following method of calibration by measurement of band swept is quite convenient to use. Radar-transmitter output should be fed to a matching load resistor or to a transmission line and antenna of the type standard for the system under test, to insure normal oscillation and modulation. Receiver input should be connected to the output of an unmodulated standard-signal generator and receiver output (from the beat-note amplifier) viewed on a cathode-ray oscilloscope with time-base sweep synchronized by the radar modulation. There will be no beat-note amplifier output except at the instants when the modulated radar-transmitter frequency sweeps through the single frequency produced by the signal generator. At such instants the amplifier will pass momentarily the beats

between the standard signal and the mixing signal fed internally from radar transmitter to receiver. These isolated beats appear as sharp spikes or "pips" on the oscilloscope trace.

Two pips appear, one during the frequency upsweep and one during the downsweep of the radar modulation, so long as the signal generator is tuned to some frequency within the band swept. As the generator is tuned toward either limit of the sweep, these pips may be seen to approach one another on the oscilloscope, finally merging into one and disappearing as the generator is tuned out of the swept band altogether. Readings are made of the signal-generator frequency when tuned so that the single coalescent pip just appears respectively at the two ends of the modulation sweep, and the sweep width is given by the difference of these two readings. Accuracy attainable by this method is limited by two factors: difficulty in judging the stage of the coalescence and disappearance at which the test-signal frequency just coincides with the frequency limit of the sweep, and inaccuracy inherent in measuring sweep width as a small difference between two large radio frequencies. The latter procedure magnifies the effect of any minute errors made in determining the radio frequencies themselves.

Reading a scale of radio frequencies to extreme accuracy may be avoided by the use of a sinusoidally amplitude-modulated test signal in the arrangement already described. Separate pips will then be visible for the beats between the sweeping radar signal and, respectively, the carrier and side-frequency components of the modulated test signal. Adjustment of both carrier and modulation frequencies of the test signal may be made so as to place the carrier pip at one end and one side-frequency pip at the other end, or alternatively to place the two side-frequency pips at the two ends, of the radar sweep. The radar sweep width is then, respectively, equal to or twice the test-signal modulation frequency. This eliminates the small-difference magnification of signal-generator errors.

Given an accurately calibrated audio-frequency oscillator, there is of course no difficulty in determining the radar modulation frequency. This may then be used with

the measured sweep width in calculating the radar range sensitivity k_R in beat-note cycles per foot of range [see equation (II.22b)]. The difficulty of accurately judging the end points of the sweep still remains a limitation, however. The variable-frequency beat method of calibration is fully applicable only to calibration of radar systems in which continuously averaging devices are used to measure range-beat frequency. Other beat-frequency measuring devices may give readings affected by details of the frequency-modulation wave form.

b. ***Time-Delay Method.*** A very direct method of calibration results from connecting between radar transmitter output and receiver input a circuit which transmits signals at all frequencies involved with a single definitely known time delay, and with a suitable and fairly constant attenuation of amplitude. This method in its most direct and least convenient form requires that transmitter and receiver each be connected through a line of known electrical length to a highly directive and properly matched antenna system. At least one well isolated radar target of very large effective echoing area must be located at a suitable and accurately known range from these antennas, in their common direction of maximum signal. Such a method has been used, for lack of available alternatives, in setting up the experimental *AN/APG-17A(XN)* equipments. Its only drawback is the practical difficulty likely to be encountered in providing sufficiently large and well isolated radar targets at desired ranges, but this disadvantage is a very serious one.

Next in order of directness is the method using a section of ordinary non-selective radio-frequency transmission line, of suitable and accurately known electrical length and suitable attenuation, connected between radar transmitter output and receiver input. Inclusion of a variable attenuator circuit in such a transmission path is often advisable. Since these transmission-delay methods produce a completely normal radar signal, they may be used not only with a range counter of known sensitivity to determine radar range sensitivity k_R but also to determine the overall range sensitivity of systems using any desired frequency-indicating device.

Delay-path length for calibration should be so chosen that modulation normal for the system under test will produce a range-beat frequency also normal for that system. This condition cannot always be met without using unreasonably bulky delay lines or encountering such excessive attenuation that no usable delayed signal results. With radar systems designed for sufficiently low radio frequency and short range, this method of calibration is entirely satisfactory. With systems designed for high radio frequency and long range, however, use of actual line sections for calibrating-delay elements becomes most unsatisfactory.

In a series of calibrating units designated *TS-10/APN*, line sections approximately 70 and 380 feet long are compactly mounted in a carrying case. Two eight-foot connecting sections of line and a variable attenuator are also provided. By use of the line sections separately and together, equivalent nominal radar ranges of 65, 297, and 350 feet are available. Exact range equivalents depend upon velocity of signal propagation in the actual cable used, and are placarded on each unit. This portable delay unit has been very useful in calibrating the low-altitude range of the **AN/APN-1* series of altimeters and their predecessors, all of which operate at a mean radio frequency of 440 megacycles per second.

Artificial delay-line structures have proved valuable in providing greater time delay with lower bulk, weight, and attenuation than is practically attainable with uniform transmission lines. Artificial delay lines composed of many cascaded low-pass filter sections are well known but are not suitable for f-m radar calibration. This is so because the very wide pass band necessary for such filters, from zero frequency to hundreds of megacycles, corresponds to an excessively small time delay per filter section.

It is also generally known that all signals having frequencies in a region for which a given circuit produces a phase shift varying linearly with frequency are delayed in time by the same amount in passing through that circuit. This common time delay τ is the slope $d\psi/d\omega$ of the linear portion of the phase characteristic of the circuit, with phase angle ψ expressed in radians. An ideal band-pass filter with matching termination has a phase characteristic

that is linear in the central region of its pass band, and provides a total phase change of π radians from one limit of the band to the other. By using a reasonably narrow pass band, a respectable time delay per section may be obtained from a multi-section band-pass filter.

High-quality filter elements are necessary in order to approach ideal-filter characteristics. Resonant circuits formed of short sections of air-dielectric coaxial line have therefore been useful as delay-line elements at ultra-high frequencies. These sections are short circuited at one end and adjustably capacitance loaded at the other, so that they operate as tunable high-Q parallel-resonant circuits. Capacitive coupling to such circuits is obtained through apertures cut in the outer conductor, opposite the capacitor plate loading the inner conductor. An element of this sort,[5] shown in Fig. VII.-5 with two oppositely directed coupling apertures, forms a complete T-type band-pass filter section, with parallel-resonant shunt arm and purely capacitive series arms. Many such

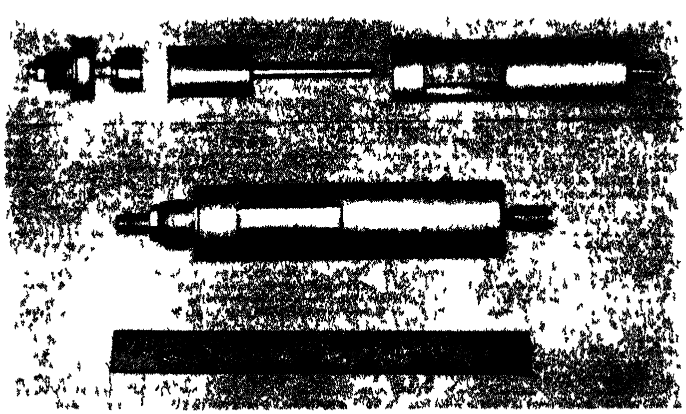


Fig. VII.-5. Filter element of artificial delay line.

sections may conveniently be mounted lengthwise between two parallel metal plates, thus being accurately positioned for stable coupling.

Using elements that are precisely constructed, pre-tuned, and mounted, a delay line thus assembled from well over one hundred elements has been found to exhibit reasonably smooth response and uniform delay over the central

portion of its pass band. An artificial delay line having an equivalent radar range of 1800 feet or time delay 3.66 microseconds over the band 445±2 megacycles per second, designated *TS-59/APN-1* and mounted in a carrying case, has been in production. It has proved very useful for calibrating the 4000-foot range of f-m radar altimeters. Its total pass-band width is 16 megacycles and overall mid-band attenuation approximately 38 decibels.

The variable-frequency beat method may be used in conjunction with f-m radar apparatus to determine quite accurately the electrical length of an unknown delay line, so that the line can then be used to calibrate radar equipment. With the unknown length of line (and an attenuator if necessary) connected from radar transmitter output to radar receiver input, an amplitude-modulated r-f signal generator is loosely coupled to the receiver input, and an oscilloscope with sweep synchronized with the radar modulation is connected to show the output from the beat-note amplifier of the radar. In the absence of signal-generator output, the oscilloscope will display stably a normal f-m radar range-signal wave of the general type shown at position (7) of section (c) of Fig. IV.-11.

Radar modulation sweep should be adjusted to the maximum available, or at least to a value giving the maximum number of beat-note cycles per unidirectional modulation sweep that can be clearly observed when a single modulation sweep (one-half modulation cycle) is expanded to the full width of the oscilloscope screen. Turning up the signal-generator output and tuning its carrier frequency to the middle of the radar band, three pips will be superimposed on the sinusoidal oscilloscope pattern. By careful adjustment of carrier frequency F_0 and modulation frequency F_m of the test signal, the two side-frequency pips can be placed at accurately similar positions on two cycles of the oscilloscope pattern which are separated by the full width of the screen, as shown by the arrows in Fig. VII.-6. The number N of whole cycles between these marker pips may then be counted. Spacing the marker pips in frequency by slightly less than the full radar sweep width avoids the obscurity of the range-beat wave form just at the points of modulation turn around and so permits adjustment

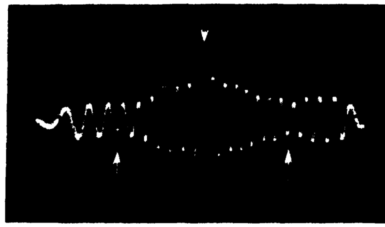


Fig. VII.-6. Radar output wave form with frequency markers for delay calibration.

of marker separation to an accurately integral number of range-beat cycles.

The upper-frequency marker has a frequency F_0+F_m and a period $1/(F_0+F_m)$, while the lower marker has a period $1/(F_0-F_m)$. The radar-signal time delay τ in the line is equal to a number N_1 (not in general integral) of periods of the upper marker frequency or a number N_2 of periods of the lower marker, so that

$$\left.\begin{aligned} N_1 &= (F_0+F_m)\,\tau \\ \text{and} \qquad N_2 &= (F_0-F_m)\,\tau \end{aligned}\right\} . \tag{VII.1}$$

In sweeping the radar signal from the upper to the lower marker frequency, the number of cycles of phase difference between the signal through the delay line and the internal mixing signal changes from N_1 to N_2 (neglecting delay in the mixing-signal path). The number N of complete beat-note cycles observed during this sweep is therefore N_1-N_2, so that

$$N = 2F_m\tau \, . \tag{VII.2}$$

Traveling freely at velocity c, the signal would in time τ cover an equivalent distance L_e, or traverse twice an equivalent radar-echo range R_e given by

$$R_e = \tfrac{1}{2}L_e = \tfrac{1}{4}Nc/F_m \, . \tag{VII.3}$$

More exactly, this is the excess of range equivalent of the external delay path over that of the internal mixing path.

c. ***Multiple-Frequency Beat Method.*** A normal radar range-beat signal is developed in the normal way when

calibrating by the time-delay method. No normal radar range output at all is developed in the variable-beat method. Still another calibrating method is known in which a seemingly normal range beat is developed entirely artificially. This may be termed a multiple-beat method of calibration, for reasons soon to appear. The frequency of the beat note so produced varies in proportion to rate of change of transmitter radio frequency, just as does the range-beat frequency in normal operation.

Each time the number of radio wave lengths contained in the total path traversed by the radar signal changes by unity as the result of transmitter-frequency shift, one complete cycle of the range beat between transmitted signal and radar return occurs. Let the total path, of length $2R$, contain N wave lengths of a signal having wave length λ_0 or frequency F_0, $N+1$ of a signal having wave length λ_1 or frequency F_1, $N+2$ of a signal having wave length λ_2, etc. Then

$$(N+1)-N = {}^{2R}/_{\lambda_1} - {}^{2R}/_{\lambda_0} = (N+2)-(N+1) = {}^{2R}/_{\lambda_2} - {}^{2R}/_{\lambda_1} = 1$$

$$= {}^{2RF_1}/_{c} - {}^{2RF_0}/_{c} = {}^{2RF_2}/_{c} - {}^{2RF_1}/_{c}\ , \text{ etc.} \qquad \text{(VII.4)}$$

Transmitter frequency shift between successive beats in normal radar operation at range R is thus seen to have the single definite value

$$\Delta F = \tfrac{1}{2}\,{}^{c}/_{R} = {}^{1}/_{\tau}\ , \qquad \text{(VII.5)}$$

where τ is the time delay in signal propagation to the target and back.

It is evident that if fixed standard frequencies are provided at uniform intervals ΔF throughout the band swept by frequency modulation of the radar, the sweeping radar signal will produce a momentary beat with each one. The frequency of occurrence of these multiple beats will be just that of the normal radar beat for a range R, providing a very useful method of calibration. The beats between the sweeping and fixed frequencies will remain only momentarily at frequencies within the pass band of an audio amplifier, so that one sharp audio pulse only will be produced for each beat. While these pulses recur at the

proper range-beat frequency, their wave form is poorly suited for counter operation. Practical utilization of the multiple-beat method of calibration therefore poses two problems: provision of the many accurately spaced individual standard frequencies required, and "smearing" of the train of individual momentary-beat pulses into a more nearly sinusoidal radar-beat-frequency output wave.

It is well known that the output of a radio-frequency oscillator amplitude modulated by a periodic train of very sharp pulses is equivalent to a large number of fixed sinusoidal component signals, having substantially equal amplitudes and separated in frequency by exactly equal intervals. The frequency difference between adjacent components is of course the repetition frequency of the pulses, while the number of equal-amplitude components is substantially twice the ratio of the time interval between successive pulses to the duration of a single pulse.

Whether positive pulse modulation (oscillator active only during pulse) or negative modulation (oscillator off only during pulse) is used makes no significant difference. The very large signal component produced in the latter case at the unmodulated oscillator frequency is of no especial use in radar calibration, and may be a nuisance. The problem of getting enough standard frequencies is therefore reduced to that of producing sufficiently sharp pulse modulation of an oscillator operated at approximately the frequency of the radar transmitter. Range simulated, determined by component-signal spacing in frequency, is controlled by the frequency of the pulse-generating oscillator and may easily be very accurately determined and maintained.

Several methods of spreading the beat wave form have been devised. A very simple method, which has probably not been tried, involves production of beat pulses at twice the required frequency and their application to control an aperiodic scale-of-two counter, or "flip-flop circuit." This then provides a square-wave output at the required frequency; the output wave can of course be made triangular if desired by simple electrical integration. Either of these wave forms is well suited for operating a limiter and averaging counter.

A very direct method of beat-wave-form smearing has been

found successful experimentally.[6] The beat output developed by a detector fed with frequency-modulated radar signal and pulsed-oscillator calibrating signal is amplified by a wide-band video amplifier, which produces a continuous beat output of varying frequency. This beat signal is then limited to remove fortuitous amplitude modulation and applied to a broad resonant circuit tuned to half the pulse frequency. The tuned circuit produces a smooth amplitude modulation of the variable-frequency video beat, with one maximum for passage of the radar frequency across each interval between adjacent pulsed-oscillator side frequencies. Rectified video output from the tuned circuit provides a suitable audio signal for radar calibration. Neither the radio frequency of the pulsed oscillator nor the resonant frequency of the tuned circuit is at all critical for proper operation, but it has been found convenient to operate the pulsed oscillator at a subharmonic of the radar frequency in order to avoid pulling either oscillator frequency by the beating process.

An especially ingenious method of beat smearing has been used in a manufactured calibrator for the **AN/APN-1* altimeter, and particularly for the high range of that equipment. This very useful device, designated *TS-250/APN*, employs a small injected signal from the radar transmitter to synchronize the starting phase of the free-running pulsed oscillator. This oscillator, tuned somewhat below the band swept by the altimeter signal, runs most of the time and is very sharply and briefly pulsed off at a pulse frequency controlled by a quartz-crystal oscillator.

Operation of the device can conveniently be explained in terms of modulated waves rather than of equivalent steady side-frequency signal components. Each burst of calibrator oscillation is of very accurately timed duration, and each produces a burst of beat signal in the upper high-frequency radio band when mixed with the radar signal in the detector of the altimeter. Each burst of beat signal starts in the same phase because of calibrator synchronization by radar-signal injection, but the very sudden termination of the burst after a fixed and definite time interval occurs at a beat-signal phase determined by the instantaneous frequency difference between the radar signal and the free-running

calibrating oscillator during the burst.

Each total beat-signal burst contains a direct-current component of magnitude depending upon the fraction by which the number of cycles in the beat-wave train departs from the nearest integer. This extra fraction of a cycle varies smoothly with the instantaneous frequency difference between altimeter and calibrator, passing through zero at successive altimeter frequencies which differ by just the pulse-repetition frequency. The output of the detector thus includes direct-current pulses, recurring at the pulse-repetition frequency of the calibrator and varying in amplitude with the instantaneous frequency difference between altimeter transmitter and fixed-frequency pulsed oscillator. The envelope of this medium-frequency train of d-c pulses is just the audio signal needed for calibration. Operation is again uncritical to pulsed-oscillator frequency, so long as this is stable and remains outside the band swept by the altimeter transmitter. In this case the frequency-pulling phenomenon is not avoided but put to good use in synchronizing the two oscillators momentarily at the end of each quenching pulse.

Side-frequency analysis of operation of the *TS-250/APN* has not been carried out, but it is evident that the sudden quenching of calibrator oscillation must cut off each beat burst sharply enough to distinguish a fraction of a cycle at ultra-high frequency. This ultra-fast modulation must correspond to the presence of uniformly strong side-frequency components throughout the frequency band of the radar. Phase modulation of the pulsed oscillator by initial synchronization to the radar signal is evidently periodic at the modulation frequency of the radar, and occurs in small steps at the pulse frequency of the calibrator. This phase modulation must spread each side frequency of the pulsed oscillator into a side band with components separated by the radar modulation frequency.

Such additional signal components no doubt account for the observed smoothing of the audio beat-envelope wave. Sharp but weak beat pulses can be observed in the audio output from the detector in addition to the desired smooth wave; these are attributed to non-synchronized transient oscillations shock excited by the modulating pulses. This

is the only one of the known methods of beat broadening that requires no detector or amplifier external to the altimeter.

To avoid disturbance of calibration by fixed error, the *TS-250/APN* equipment is arranged to frequency modulate its pulsed oscillator slightly at a sub-audio frequency. This destroys phase coherence of the beat signal on successive modulation cycles of the altimeter and so permits the altimeter counters to average out fixed error. Accurate audio frequencies for range-counter calibration are also provided by this test unit, as well as a calibrated attenuator for checking overall sensitivity of the altimeter. A device of the same type was expected to make possible accurate calibration of the *AN/APG-17A* equipment, but never became available.

d. ***Range-Counter Calibration.*** A simple, continuously acting range counter in proper working order is not subject to obscure errors, especially if linearized to average out speed frequencies or not fed with signals affected by speed. Such a counter may be calibrated simply by use of an oscillator capable of providing two accurately known steady test frequencies, one low and the other high in the operating range of the counter. Calibration both of zero and slope of the counter characteristic may be made by applying the two frequencies alternately and adjusting alternately the counter sensitivity h_R (capacitor or limiter-output swing) and the bias voltage to which the counter load is returned. Sensitivity is to be adjusted at the high frequency and bias at the low frequency. This process of successive approximation is continued until the output indicator with which the counter normally operates gives exactly its intended indications for both test frequencies. Care must be taken to use a sufficiently accurate calibrating oscillator, as not all oscillators are themselves calibrated accurately enough for this use.

In the case of range-only devices like the **AN/APN-1*, both indicator and limit counters are calibrated as above; the indicator counter is calibrated first and with its indicator is then used with a variable test oscillator to center limit-circuit operation properly at two check points. Overall calibration then requires introduction of a real or

simulated time delay (such as *TS-10/APN*, *TS-59/APN-1*, or *TS-250/APN*) between transmitter output and receiver input. Either modulating frequency or width of modulation sweep is next adjusted to calibrate radar range sensitivity, until the range indication at the counter output has the value proper for the time delay introduced (491.6 feet of range per microsecond delay). Due allowance must be made for any range residual which it is not desired to indicate. In the absence of residual range or extra counts, single-frequency calibration of sensitivity only is sufficient.

In the case of range sensitivity of switched counters, simple calibration with a test oscillator should not be relied upon. Switch timing can introduce serious error into such a calibration (see section 3d of Chapter IV.). In this case, simulated radar signals for two ranges are required for overall calibration in range; they are to be used alternately like the two test-oscillator signals above, adjusting alternately overall range sensitivity (sweep width) and counter-load bias until correct indication is given of both ranges.

4. SPEED CALIBRATION

a. *Speed Calibration of Counters.* In the case of devices like the *AN/SPN-2* using basically a simple averaging counter to measure speed only, calibration can be established, as in the case of the simple range counter, by using one or two fixed test frequencies in the working range of the counter. Overall calibration of speed-only devices is accomplished by first calibrating the counter sensitivity h_s alone with a test oscillator as above. By then measuring the radio operating frequency F_0 in megacycles, radar speed sensitivity k_s in cycles per second per foot per second may be calculated simply by division of F_0 by 491.6. In more complicated cases, however, it is necessary for calibration to simulate the effect of speed on an f-m radar range signal.

Frequency variation of the beat-note output of an f-m radar, shown in Fig. IV.-13, may be imitated by switching at the radar modulation frequency between the outputs of two test oscillators. One test oscillator must work at range-beat plus speed-beat and the other at range-beat

minus speed-beat frequency. Such a switched-oscillator arrangement may be used to determine or adjust both range and speed sensitivities of a switched counter when the oscillator and counter switching cycles are synchronous. Effective switched-counter speed sensitivity depends, however, on the sweep-reversal phase lag (see section 3d of Chapter IV.) of the radar with which the counter is to operate. The oscillator switching used in calibrating the counter must therefore be made to lag the counter switching in phase by the same amount as does the radar modulation. Radar-modulation phase lag is rather difficult to measure accurately, thus setting a practical limit to the accuracy of switched-counter calibration with a switched pair of oscillators.

Phase jumps at switching between two free-running oscillators of different frequencies are quite different from those in a real f-m radar signal, so may in some cases give rise to different fixed-error effects. To avoid this, a calibrating wave form like the actual radar-beat wave form has been produced with a beat-frequency test oscillator. This is accomplished by switching the tuning of one of the beating oscillators, in synchronism with radar modulation, between two discrete values producing frequencies different by twice the range-beat frequency to be simulated. The other of the beating oscillators is fixedly tuned to a point, between the two switched-oscillator frequencies, so chosen that beat frequencies $f_R + f_S$ and $f_R - f_S$ are produced alternately.

No voltage discontinuity nor even marked phase discontinuity occurs in the output of the first beating oscillator when its tuning is switched, nor do such discontinuties occur in actual r-f beating signals of f-m radar. The further fact that upper-frequency and lower-frequency oscillators exchange places upon switching, just as do transmitted and radar-returned signals upon modulation turn around in actual radar ranging, results in a beat-frequency oscillator signal having just the sort of wave form exhibited by actual radar signals. In this case also, oscillator switching must be delayed to simulate radar modulator lag. The usual drift and accuracy difficulties of beat-frequency oscillators are present and have prevented this method of

signal simulation from becoming of great practical value.

b. *Simulation of Radar Speed Signal.* Oscillator-switching methods for switched-counter calibration, such as have just been described, are rather cumbersome and offer various possibilities for the occurrence of hidden errors in overall radar-system calibration. More direct simulation of received radar signals combining effects of speed and range is desirable. The effect of constant speed of closing of the radar on the target is, as described in section 2c of Chapter II., a substantially constant increase in frequency of the received signal over that of the transmitted signal, and is called Doppler effect. This increase in frequency may also be regarded as a steadily increasing phase advancement of the received signal with respect to the transmitted signal. The steady phase advancement occurs at a rate which is proportional to speed and which often amounts to many complete radio-frequency cycles per second.

When a delay path, real or artificial, is connected between the radar transmitter and receiver to simulate the effect of actual target range, the effect of speed may be simulated by introducing into that path a phase shifter which operates to advance continuously and without limit the phase of the delayed signal. At low frequencies, such a continuously advancing phase would be produced for example in the rotor output of a synchro unit having its stator excited by polyphase supply, with the rotor driven by a motor at the rate of one revolution per complete cycle of phase advance required. At the ultra-high radio frequencies used in radar, the basic requirement is the same but a very different phase-shifter structure is necessary.

Let radar transmitter and receiver be connected to two perfectly terminated transmission lines, so that radio-frequency energy applied to the near end of either line will be propagated uniformly toward the far termination, with no energy returned and in consequence no standing waves on the lines. Let these lines be placed parallel to one another, with feed ends adjacent, and a slidable connection be made between them, as in Fig. VII.-7(a). As this connection *B* is slid toward the feed end of the lines, the phase of the output from line *C* will be advanced, and

indeed will advance continuously by one full cycle per half wave length of displacement of B (the displacement being effectively doubled because made along both lines at once). Total phase shift practically procurable with straight lines is of course strictly limited.

The limitation may be removed by forming the two lines into circles in parallel planes, as in Fig. VII.-7(b). If

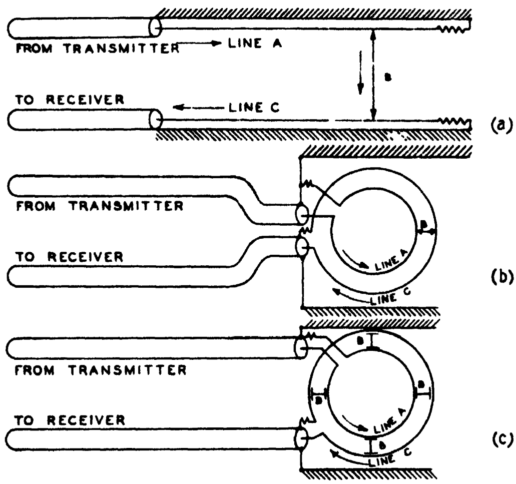


Fig. VII.-7. Basis of operation of continuous phase shifter for radio frequencies.

the circle circumference is exactly a whole number of wave lengths, the phase of the traveling wave on each line will be exactly the same at both ends of the small feeding gap in the circle. Except for the slight amplitude difference caused by attenuation in the circle of line, there will then be no change in output as the connecting bar crosses the gap between beginning and end of each circle. Feeding and terminating points of each circle must of course be isolated, by breaking the circle, to maintain the condition of unidirectional energy flow and uniform phase progression along the lines. By moving the connector steadily and repeatedly around the transmission-line circles toward the transmitter, a steady advance of output phase may be produced and, in effect, the output signal will have a frequency higher than that of the input signal.

A number of practical points may be noted. Conductive

sliding connections are noisy and unreliable, therefore a capacitive connection provided by a hole in a rotatable metallic shielding disc between the circular lines is preferred. Cross connections made one-half wave length apart provide total paths different by just one wave length and therefore produce in-phase signal components which reinforce output. To minimize disturbance on crossing the gap in each circle, these gaps in the two lines may be separated circumferentially. These modifications are indicated in Fig. VII.-7(c), which is drawn for circles just two wave lengths in circumference and shows four capacitive cross connections *B*.

For lines in free space, the connector must move just as far as does the target in the actual radar case to produce the same phase advance. That is, the connector must actually move with the linear speed which it is to simulate. By embedding the phase-shifting lines in the faces of parallel-mounted discs of high-dielectric-constant ceramic material, velocity of propagation on the lines is greatly reduced, permitting a phase shifter of small dimensions to be made and reducing proportionately the speed at which the connector holes must be moved. The "inner" conductors of the lines are metal filaments embedded in the ceramic-disc faces and their "outer" conductors are metal plates which back and support the ceramic discs. There is always some difficulty in maintaining acceptably small the phase discontinuity at the line gaps and the residual standing wave on the circular lines. These tolerance difficulties are enhanced by the use of small high-dielectric lines.

Continuous phase shifters of this sort have been used very successfully at 410 and 515 megacycles per second for calibrating *AN/APG-4* and *AN/APG-6(XN)* equipments. Adaptation to 1500 megacycles has been less satisfactory, though usable results have been attained with lines of low dielectric constant. With high-dielectric lines containing a large number of wave lengths, maintenance of integral-wave-length circumference and consequently of acceptable phase discontinuity at the gap has been excessively difficult at 1500 megacycles. For speed calibration, the phase shifter is driven by a d-c shunt motor with coarse and fine

speed controls; the simulated speed is indicated by a directly connected aircraft-engine tachometer, with dial suitably calibrated in speed units.

That type of multiple-beat range calibrator in which starting of a pulse-quenched oscillator is synchronized with the transmitted-signal phase also admits speed simulation by means of a continuous phase shifter. In this case, the phase shift must be applied in the injection path for the synchronizing, so that starting phase of the pulsed oscillator will steadily advance with respect to the phase of the mixing signal fed from transmitter to receiver within the radar. In the case of multiple-beat arrangements using an altogether free-running pulsed oscillator, the phase shifter provides no simulation of speed.

Simulated speed can also be provided when using a range calibrator with a free-running pulsed oscillator, but an entirely different approach is required. Range is simulated by the momentary beats resulting as the radar transmitter frequency sweeps back and forth across a "picket fence" of equally spaced fixed side frequencies produced by pulse modulation of the calibrating oscillator. If these side frequencies are not fixed but steadily increasing, the frequency of occurrence of beats is reduced during upsweep and increased during downsweep of the radar modulation, as the radar transmitter frequency changes respectively with and against the changing calibrator frequencies. Steady increase of the calibration frequencies is of course not practicable, but an unsymmetrical-sawtooth frequency modulation of the pulsed oscillator will produce periods of steady increase of frequency, each followed by a very rapid return to an initial frequency and then by another steady increase, and so on.

Speed simulated may be determined in the above case by counting beats between the frequency-modulated pulsed oscillator and a fixed-frequency speed-reference oscillator, as the "picket-fence" side frequencies of the former move steadily past the fixed reference. Those jumps caused by the sawtooth nature of the calibrator frequency modulation may be minimized by artifices of modulation control. This type of combined speed and range simulation has been found operative by experiment, but has not been put to use.

5. Overall Calibration Procedures

a. *AN/APG-4.* When calibrating a radar system that uses a switched counter sensitive both to range and speed, and operating under conditions dependent upon both range and speed, it has proved advisable to use an overall calibration procedure with a radio-frequency test signal capable of simulating the effects of both variables at once. A simulator to produce such a signal, designated *TS-51/APG-4*, has been manufactured for the testing of Sniffer equipments; this device is shown in Fig. VII.-8. It includes a motor-driven phase shifter of the sort described in section 4b above, with speed controls and tachometer, in

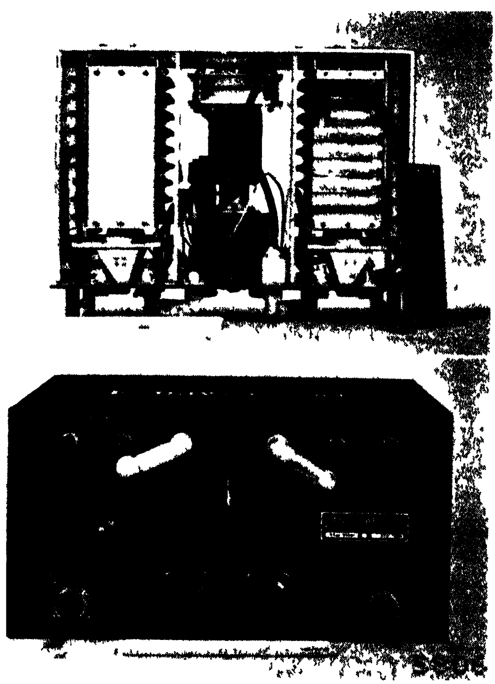


Fig. VII.-8. Complete range and speed simulating calibrator *TS-51/APG-4*.

addition to two separate sections of artificial delay line made of many band-pass elements as described in section 3b, and a lamp to indicate occurrence of release. In the figure, the two artificial lines may be seen in the side compartments and the assembly of tachometer, motor and phase shifter in the central compartment. The line-

carrying ceramic discs of the phase shifter are visible at the rear of the center compartment, with a thin rotatable shutter disc between them to provide the moving connections. Exact range equivalent of each line is marked on the individual test set.

Data on time lag in release and on residual range in r-f lines is required for each individual type of Sniffer installation, so any single calibration can only be correct for identical installations. Equivalent release ranges for calibration are determined by subtracting residual range R_r from placarded range equivalents of the artificial lines, singly and in combination. In terms of speed limits and overall time lag for which any particular altitude compensation unit is designed, values of approximation slope T' and intercept S_0 may be determined readily for several representative altitudes within the range of the compensation unit, by use of equations (V.23) and (V.28) subject to (V.24). Speeds producing release at the calibrating ranges may be determined and tabulated for the chosen altitudes, either from linear range-speed graphs drawn with the proper slopes and intercepts or by simple calculation from equation (V.12).

Calibration of the *AN/APG-4* bombing radar with the help of the *TS-51/APG-4* test set is a three-way process of successive approximation. The phase shifter and one or both delay lines are connected in cascade from transmitter output to receiver input receptacles of the equipment under test. Compensation of the *AN/APG-4* system for a chosen altitude is provided by setting the *SA-9A/APG-4* or *SA-28/APG* compensation unit. The speed at which the phase shifter is driven is then slowly raised until the release relay closes, and the release speed noted. An adjustment is made, then a release trial under new conditions, then a different adjustment, and so on.

Modulation-amplifier gain and thereby sweep width is set with low-altitude compensation and long-range delay line for release at the proper high speed. Bias voltage to counter load is set with low-altitude compensation and short-range delay for release at the proper low speed. Form of altitude compensation is set by adjusting the total value of resistor r_2 in the compensating network

(see Fig. VI.-30). This is done with high-altitude compensation setting and long-range delay, to give release at the proper low speed as predetermined from the release approximation. The three adjustments interact, so correct values can only be reached by orderly cycles of repeated adjustment. Only these three calibration controls are available in the *AN/APG-4* or *AN/APG-6*, so when they are set release must be acceptably accurate under all other altitude-range-speed conditions without further adjustment.

b. *AN/APG-17A*. The 1500-megacycle equivalent of a combination of *TS-51/APG-4* and *TS-250/APN-1*, designated *TS-404/APG*, was to have been used in calibrating *AN/APG-17* and *-17A*, but was never completed. Procedure for its use with *AN/APG-17* would be essentially the same as that described above for calibration of *AN/APG-4* with *TS-51/APG*. Greater flexibility of the *AN/APG-17A*, however, requires a different calibrating procedure. The calibrating controls are greater in number but ineraction among them is almost entirely avoided. A precise variable voltage divider, graduated in per cent of cathode-follower grid voltage at release, is required in addition to the *TS-404/APG* and an audio-frequency oscillator to calibrate the *AN/APG-17A*; this divider is fed from the plate supply of the equipment under test.

Feeding the counter-load resistor from the precision divider and with counters disabled, relay-tube bias is adjusted to cause release to occur at the 100-per cent setting of the divider. With speed counter disabled and range counter acting, a convenient audio frequency is then fed to the counter system and the range-counter capacitor adjusted to reduce the bias-divider setting for release by roughly 12 per cent per kilocycle of the frequency in use. With audio frequency set for release at a chosen calibrating-divider reading near the center of the normal bomb-channel bias range, counter load is returned to the normal bomb-channel bias control. The bias-scale (bias-divider current) adjustment of the *AN/APG-17A* is then set so that release occurs at a percentage reading of the bomb-channel bias dial which is equal to the chosen release reading of the calibrating divider. The *TS-404/APG* is next connected between transmitter output and receiver

input to simulate radar range of an amount placarded on the test set. Altitude-compensation factor F and range-scale factor R_0 of the equipment under test are adjusted to any convenient values which together will give a mid-speed release range R_0/F of the value placarded. With speed counter still disabled, modulation-amplifier gain is set to give a radar range-beat frequency of approximately 6.0 kilocycles, as indicated by release at a bias reading of 28 per cent.

The range counter is then disabled and speed counter activated. Speed-counter capacitor balance is adjusted until bias for release is not affected by varying the radar sweep, and thereby the range-beat frequency, in the vicinity of the value for mid-speed release. Balancing of the biases applied from the cathode follower to the return leads of the switched counters is then accomplished by varying the tapping point at which the positive-counter discharge diode is connected to the cathode-follower load resistor. The tap is set to the point for which counter-capacitor balance is unaffected by changing the counter-input voltage swing; this adjustment equalizes the total effective voltage swings applied to the positive and negative switched counters [see equation (IV.6)].

The speed counter being thus fully balanced, the scale of the speed-sensitivity control is next calibrated, still in the presence of normal range signal. The bomb-channel speed-sensitivity control is first set to a convenient value, for example 0.25 per cent per knot. The change of counter output for release upon changing by a known amount the speed simulated by the motor-driven phase shifter is then observed. Finally, the plate-current swing of the limiter is adjusted to cause the output change observed for the speed change made to correspond to the speed-sensitivity scale reading.

The speed counter is then again disabled and the range counter activated, so that the range-counter capacitor may be set finally to give a counter sensitivity of just 12 per cent per kilocycle per second. Final adjustment of sweep width to give a radar range sensitivity of just $6.00F/R_0$ kilocycles per second per foot is also required. Bias scale and speed-counter bias balance are not likely to

require major readjustment in equipment previously calibrated, nor need the preliminary settings of range-counter and radar-range sensitivities be made in such cases. Overall checks of range-speed-altitude relations at release under a number of selected conditions are of course to be made as a final precaution against accumulated error.

These two examples of overall f-m radar calibration, by successive approximation in a specialized system and by a lengthy but straightforward procedure in a relatively flexible system, should illustrate adequately the methods to be used in meeting exact operating requirements. Some general considerations should also be mentioned, however. Availability of a continuous phase shifter for speed calibration permits its use at very low running speeds as a convenient means of averaging out fixed error in range-only calibration. In speed-only or range and speed checks, care must be taken to avoid simple commensurability between range and speed frequencies with consequent return of fixed error.

The procedure of fixing range and varying speed for release when calibrating is, of course, a reversal of the normal airborne operating conditions. It is a natural consequence of the use of physical delay lines of inherently fixed length for calibration, and seems to produce no harmful result. Range could conveniently be made variable in the multiple-beat calibrators, but provision of a useful set of accurately fixed simulated speeds for use with variable range is not convenient. Even with fixed speed and variable range, calibration would be a static process whereas actual operation is dynamic. The question of whether static calibration is altogether adequate has not been fully investigated. Some attempts at dynamic calibration, however, have clearly established that the static method requires simpler equipment and is easier to use.

6. NOTATION AND REFERENCES

a. *Notation.* The notation listed alphabetically below has been used in this chapter.

- c Velocity of propagation of radio waves in sea-level air.
- C Circuit capacitor, usually with identifying subscript.

F Radio frequency, with identifying subscript if under special condition; also, general altitude-compensation function.

F_0 Mean radio frequency of modulated wave; also, frequency of variable-beat calibrating oscillator.

F_m Frequency of modulation of variable-beat calibrating oscillator.

ΔF Change of radar frequency to change standing-wave pattern by one full wave.

L_e Electrical equivalent length of delay line.

N Number of standing waves between radar and target, with identifying subscript if for special condition.

r Circuit resistor, usually with identifying subscript.

R Range or distance between radar and target.

R_0 Reference range for level-flight rocket firing.

R_e Electrical range equivalent of delay line.

R_r Residual range equivalent due to propagation delay within equipment.

S_0 Speed-axis intercept of straight-line range-speed approximation for bombing.

T' Slope of linear range-speed approximation for bombing.

V Vacuum tube, usually with identifying subscript.

λ Wave length of radio wave, with identifying subscript if for special condition.

λ_0 Length of radio wave of frequency F_0.

τ Time interval required for wave propagation.

ψ Phase angle.

ω Radian frequency $2\pi F$.

b. *References.*

1. D. Blitz: "Pilot Warning Circuits for the Sniffer," Report no. 21 under Contract NXsa-35042 (March 3, 1945).

2. D. G. C. Luck: "Preliminary Tests of Anti-Fade Accessory for Sniffer," Report no. 1 under Contract NXsa-35042 (February 18, 1944).

3. D. G. C. Luck: "An Anti-Fading Accessory or Short-Time Memory for Sniffer Equipments," Report no. 3 under Contract NXsa-35042 (May 5, 1944).

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5. A. J. Neuman: "An U-H-F Band-Pass Filter Delay Line," Report TR-17-C of Victor Division, Radio Corp. of America, Camden, N. J. (May 3, 1944).

6. E. O. Keizer and H. G. Fisher: "F-M Radar Calibrator," Report no. 24 under Contract NXsa-35042 (September 5, 1945).

CHAPTER VIII.

DEVELOPMENTAL SINGLE-TARGET SYSTEMS

1. General

Various other forms and applications of single-target f-m radar have been proposed from time to time, in addition to those that led to the fully developed equipment described in Chapters VI. and VII. Some special uses and equipments will be described in this chapter. While some proposed systems were not carried beyond the planning stage, a number have been at least partially built and laboratory tested, and a few were carried to the point of limited flight testing of a whole system or at least of some fundamental elements. There is, however, no doubt that only a small sector of the field of possible applications is represented here. Whether any of the systems here described are ever used or not, this discussion of uses already suggested and of ways of adapting f-m radar techniques to varied applications may serve to stimulate invention of further ways of using frequency-modulated radar.

2. Range Only Rocket Firing

The investigation of level-flight rocket firing which led to development of the *AN/APG-17A* equipment also brought to light a simple special case. For certain fast rockets fired from certain slow aircraft in level flight, slant range for firing is found to pass through a minimum as air speed of firing craft varies. This behavior is evident in Fig. V.-11(a), which is typical of an aircraft-rocket combination which happens to exhibit a minimum of required firing range at a normal operating air speed. Over a reasonable range of air speeds near that for minimum range, firing at a constant range independently of closing speed gives as good average accuracy as does any linear range-speed relation. This of course neglects wind and target speed, which in some cases are insufficient to produce much error anyway. Compensation of firing range

for flight altitude and rocket-propellant temperature of course remains necessary.

Standard *AN/APG-4* equipment was modified, in order to test this method of firing, so as to use only a simple unswitched range counter. A ballistic-correction control for impact adjustment to suit propellant temperature was used to vary the release bias. Modulation sweep width and voltage to ballistic-correction control were compensated in corresponding fashion for altitude, in accordance with the requirements of one type of aircraft and one type of rocket, but no other altitude compensation of bias was used or needed.

Limited flight tests and rocket firings with this very simple equipment indicated that the results were approximately as expected. Inadequate technique for measuring rocket impacts prevented the tests from yielding conclusive data on accuracy of firing, however. Subsequent work in development of the more complex but more generally useful *AN/APG-17A* equipment prevented further tests of control of level-flight rocket firing on the basis of range only.

3. ROCKET SIGHTING

a. ***Problem and Proposed Solution.*** There has been a need for automatic equipment to determine the angular depression of line of sight required for visual firing of rockets at surface targets from diving aircraft, and to adjust an optical sight to provide this depression. Sights are available which include a servo mechanism to set the depression angle to a value proportional to a voltage supplied from a low-impedance direct-current control source; these sights also include a graduated manual adjustment for setting the reference line from which the servo-controlled depression is measured.

It was pointed out in section 7 of Chapter V. that rocket ballistic data has been found empirically to be reasonably well fitted over a limited region by use of a sight-depression angle β related to slant range R and slant closing speed S thus:

$$\beta - \beta_0 = D(KR - S + S_0) \ . \qquad \text{(VIII.1)}$$

Coefficients D and K (which is B/D in the notation of

Chapter V., section 7) depend upon both dive angle and propellant temperature, K at least in rather complicated fashion. Sighting-reference angle β_0 and speed intercept S_0 do not depend upon dive angle or temperature. All four parameters depend on type of aircraft and rocket, though this dependence has not been adequately investigated.

The quantity in parentheses on the right of equation (VIII.1) will be recognized as the typical total output of a frequency-modulated radar using unequal switched counters. K is then the ratio of overall range sensitivity $k_R h_R$ to overall speed sensitivity $k_S h_S$, and S_0 is the ratio to overall speed sensitivity of bias voltage e_0 applied to the counter load. The quantity $\beta - \beta_0$ is proportional to the voltage input required to set to depression angle β a servo sight which has been adjusted manually for reference depression angle β_0. Coefficient D is therefore proportional to the fraction of total radar-output voltage that must be applied at low impedance to control the servo sight, in order that the desired automatic sight-depression setting may take place. Required values of β_0, S_0, K and D were found from the ballistic data to be of physically realizable magnitude in the few cases studied. Attainment of the necessary K values involves inconveniently large ratios of switched-counter speed sensitivity to range sensitivity, as well as rather small modulation sweep if range-beat frequencies are to be of usual magnitude.

An *AN/APG-4* bombing equipment was set up with appropriate counter-sensitivity ratio and suitable adjustable values of sweep width and of bias to counter load. The bomb-release relay tube was omitted and an adjustable fraction of the cathode-follower output was applied to the grid of a second cathode follower, which had sufficiently low output impedance and the proper output voltage for direct connection of its cathode to the control input of the servo sight. β_0 and S_0 were of course to be manually adjusted to suit the aircraft and rocket used. Manual adjustment of K and D for propellant temperature, as well as for aircraft and rocket, was also expected. Manual adjustment of K and D for dive angle ϕ used in attack was to be made initially, but ultimate adaptation to automatic adjustment for dive angle was desired. The final result

of section 7 of Chapter V. offers considerable hope for automatic adjustment of voltage ratio D for angle ϕ, but is not encouraging with regard to range/speed sensitivity ratio K. Counter-output voltage was expected to vary considerably during the approach, the servo sight following this variation.

b. ***Experimental Results.*** The system comprising the servo sight and *AN/APG-4* modified as above was tested extensively in the laboratory. Counter-tube bias appears between counter output to cathode-follower grid and counter-return connections to cathode-follower cathode circuit. This bias was found to vary appreciably as total counter-output voltage varied over its required operating range, and indeed the biases vary in opposite directions for the upsweep and downsweep counters. Counter-tube bias affects the effective counter-input voltage swing (see section 2c of Chapter IV.), and a differential bias variation on the upsweep and downsweep counters therefore affects their relative sensitivities.

Desired range sensitivity of the rocket-sighting counter system represents a slight difference between the large individual sensitivities of the upsweep and downsweep counters. Small fractional changes in the individual sensitivities therefore produce major changes in net range sensitivity. This effect was found to be very serious in the experimental rocket-sighting system. It may be regarded as resulting from the slight defect in following which is inherent in the counter-linearizing cathode follower. Attempts to improve follower action succeeded in reducing the residual defect, but not to an acceptable level.

Automatic control of sight setting by radar speed and range was of course obtained without difficulty. Adequate accuracy throughout the operating range was not obtained because of the counter behavior described, so no flight tests were made of the complete automatic sight-setting system. A few test flights including diving approaches to surface targets were made in order to observe radar-signal quality. Substantial reduction of maximum usable radar range with increasing dive angle was observed, even though

the tests happened to be made over unusually smooth water. It was felt that worse range reduction might have been expected if diving flight tests had been made in the presence of strong sea return from rough water.

Because of these difficulties, and because of the promising results of early study of level-flight automatic firing, further development of automatic sight setting by f-m radar for diving flight was suspended. The work done did make it quite clear that the Sniffer principle of obtaining both range and speed information from a pair of switched counters is only satisfactory for operation under the normal Sniffer condition, which is that accurate information is needed only at a single output voltage. Preliminary tests indicated that satisfactory sight-setting operation could be obtained in the laboratory by applying a servo-controlled follow-up voltage from the sight to the counter load, so as to provide null-type operation with the cathode-follower grid always maintained at a fixed voltage. Null-setting servos applied separately to range-only and speed-only counters, and jointly controlling in turn the sight servo, offer greater flexibility in adjustment of characteristics. Work along these lines was suspended before any conclusive result was reached.

4. GLIDE PATH

An aircraft must sometimes be brought down under radio guidance to make contact with the surface of the earth at a specified point. This may be necessary in order to collide with a surface target or in order to make a spot landing. Lateral guidance in such a maneuver may be given by azimuth-controlling equipment like the *AN/APG-6*. Vertical guidance may in principle be given by application of the same direction-finding techniques to determination of vertical sighting angle. Unless an extremely sharp beam of radiation is available, however, accurate vertical direction finding is always a very difficult process technically. Even with accurate angle measurement, control of aircraft position by angle data alone is a somewhat roundabout process.

Control of aircraft position in the vertical plane during a collision approach may on the other hand be quite

straightforward and accurate if done on a basis of range measurement. F-m radar lends itself particularly well to fully automatic flight control on a distance-measurement basis. The radar altimeter is able to fly an aircraft automatically at any altitude for which it has been set by the adjustment of a voltage divider, as described in section 2c of Chapter VI. A slant-range measuring radar is able to operate a servo which sets the altitude-controlling voltage divider to a value determined by the slant range to the collision target.

A straight-line glide to contact at the target results if altitude is made to decrease linearly with decreasing range, in such fashion that range and altitude reach zero simultaneously. This means simply that altitude must be maintained in a constant ratio to range, the ratio determining the slope of the straight-line glide path. Using suitably non-linear control of altitude by range, the glide path may be made to curve either upward or downward in any desired shape. Whatever the predetermined shape of the glide path, the position control of the aircraft at each point of its approach is entirely definite.

Apparatus required to set up an approach path in this way is very simple, as may be seen from the functional block diagram of Fig. VIII.-1. A normal radar altimeter

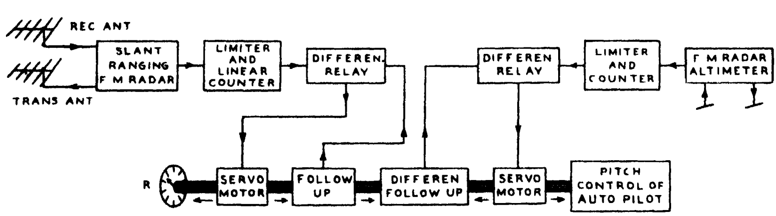

Fig. VIII.-1. Block diagram of radar glide path.

arranged for normal *ACE* operation of the aircraft, through attitude-control unit and auto pilot, is necessary. A slant-range measuring radar, for example a Sniffer with linear range counter only, is also necessary. A simple servo follower and control, similar to the *SA-28/APG*, completes the system; it is connected to produce a servo-shaft setting proportional to range output of the Sniffer, which in turn sets the altitude at which the altimeter holds the aircraft. Of course, the range servo may also

be made to modify the range sensitivity of the Sniffer to compress the required audio band. The shape of the path established is controlled by the resistance-rotation characteristics of the follow-up and control potentiometers driven by the servo motor.

The system shown was not tried as a whole. Experience with its various parts, however, indicates that it might be expected to operate well without extensive development.

5. TARGET TRACKING

a. ***General.*** Tracking of targets in such a manner as to make possible prediction of their future motion is a standard procedure in fire control, whether the necessary data is gathered by radar or otherwise. Frequency-modulated radar, with its ability to measure directly both target range and rate of change of range, and its ready applicability as a means of control, is uniquely suitable as a data source for predicting trackers.

In most trackers, some element such as a voltage, phase, or angular position of a shaft is made to vary in proportion to target range. The time rate at which variation in range occurs is of course dependent on the speed of closing of the observing craft upon the target. A rate-controlling element, or variable-speed drive, is necessary to establish a variation of the tracking element which truly represents the changing range. With most methods of collecting data, the variable-speed drive is adjusted in accordance with information derived in some way from a timed sequence of observed values of range. With f-m radar, the slow and often insufficiently accurate process of establishing a rate from successive observations is unnecessary, as closing-speed data is available directly. F-m radar is therefore especially well able to control the usual variable-speed type of range tracker expeditiously and accurately.

Trackers not requiring the complication of a variable-speed drive are possible but appear not to be widely known. These constant-speed trackers do not follow and predict range to target, but rather the interval of time which must elapse before the target is reached. If at a given instant it can be determined in some way that the target will be reached in just 10.0 seconds, then after the lapse of

5.0 seconds it can safely be predicted that 5.0 seconds have yet to elapse before the target is reached. This is always true, no matter what the speed of approach or the range at which the original time determination is made, so long as the approach speed remains constant. The predicting tracker required for time-to-target operation is merely an ordinary clock having sufficient short-period accuracy and a sufficiently open time scale. Observational data is used initially to set the zero of the clock scale so as to give a correct reading of time to target. Thereafter, observed data serves only as a check on the running of the clock, making subsequent minor corrections in setting if necessary.

The possible field of use of the time method of tracking has not been fully explored. It is clear, however, that once an accurate clock setting is established, loss of further data will not impair operation unless approach speed subsequently changes. In the case of aircraft flying over surface targets, the time predicted may be that at which the aircraft will be directly over the target; the utility of a time tracker of this particular sort in controlling bomb release is evident, as it is only necessary for release to anticipate arrival by the time of fall of the bomb. Such a tracker has often been referred to as a "long-time memory", because it can function on old data, and with a disappearing target such as a submerging submarine, the advantage of being able to release bombs on "remembered" data is readily apparent. As another use, the instant of arrival of an aircraft approaching a landing on a carrier vessel might be predicted, and this prediction might be used to synchronize the approach with the rise and fall of the flight deck as the vessel pitches.

Time tracking would involve excessive and unwarranted complication with many methods of obtaining data. Frequency-modulated radar data, however, leads to a remarkably simple system. The remainder of this section will discuss various aspects of such a system. Applications of *f*-m radar to more conventional trackers are rather obvious and no discussion of them seems necessary.

b. *Basic Time Tracker.* The relation between target range R, (constant) closing speed S, and time T required

to reach the target, when there is no relative target motion across the line of sight, is simply

$$R/T - S = 0 \ . \qquad \text{(VIII.2)}$$

The total counter-output voltage e, from a switched-counter f-m radar with its counter load returned to a bias voltage e_0, is given by

$$k_R h_R R - k_S h_S S = e - e_0 \ . \qquad \text{(VIII.3)}$$

This voltage can be made to cause a servo motor to run in one direction if e exceeds e_0, or in the other direction if e_0 exceeds e. The radar can solve equation (VIII.2) automatically if the motor is made to adjust the modulation sweep. The motor will then hold e equal to e_0 by setting radar range sensitivity k_R so that

$$\frac{k_S h_S}{k_R h_R} = \frac{1}{2f_m} \cdot \frac{h_S}{h_R} \cdot \frac{F_0}{W} = T \ , \qquad \text{(VIII.4)}$$

where F_0 is fixed radio carrier frequency, W is adjustable frequency band swept, f_m is fixed modulation frequency and h_R and h_S are fixed range and speed sensitivities of the switched counters. These results of course follow immediately from the discussion in Chapters II., IV., and V.

This scheme will work between a minimum sweep W_{min}, determined by errors caused by fortuitous residual frequency modulation, and a maximum sweep W_{max} determined by linear modulation capability of the radar transmitter and modulator. Correspondingly, it will determine time to target from a maximum value T_{max}, found by using W_{min} in equation (VIII.4), to a minimum T_{min} found by using W_{max} in that equation. Referring to the sweep-width controlling circuits of Fig. III.-12 and to (VIII.4),

$$W/W_{max} = T_{min}/T = r_2/(r_1 + r_2) \ . \qquad \text{(VIII.5)}$$

Thus if (with modulating switch in the dotted location of Fig. III.-12) r_2 is fixed and r_1 is made a linear rheostat on the servo-motor shaft, shaft rotation will directly represent time to target T on the basis that

$$T/T_{min} = (r_1 + r_2)/r_2 \ . \qquad \text{(VIII.6)}$$

In this way, the servo-controlled f-m radar provides

automatically an output shaft that is directly useful for setting a tracking clock. A motor governed to accurately constant speed makes a suitable clock. Differential gearing between the clock and any device using its time-to-target output provides a suitable way of introducing the radar-determined setting of the time-scale zero.

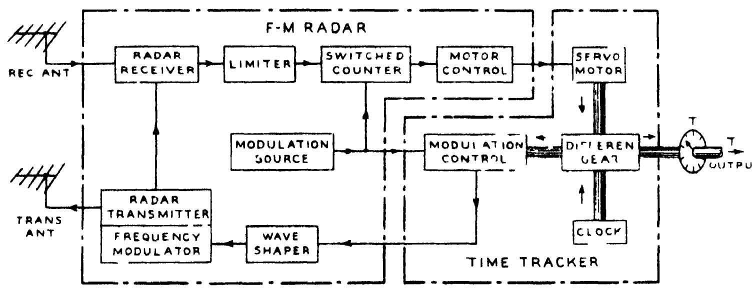

Fig. VIII.-2. Block diagram of basic time tracker.

Fig. VIII.-2 is a functional block diagram of the basic time tracker that results from the foregoing approach to the tracking problem. The extreme simplicity of the external additions required to a radar of the Snifler type may be noted. Only two minor internal modifications are necessary. One is rearrangement of the relay-amplifier circuit to fit it for controlling a motor without disturbing the action of the cathode-follower used to linearize the counters; the other is rearrangement of the modulation-generating circuit, by moving the square-wave switch to the location shown dotted in Fig. III.-12, to avoid deformation of control characteristic by extraneous loading.

Range-beat frequency does not vary with time to target in the case of the tracker shown, since increase of sweep width just keeps pace with decrease of range in the approach. Range frequency does vary with speed, of course. This arrangement permits a relatively narrow audio-amplifier pass band and consequently a good signal/noise ratio, particularly if the range of speeds to be covered is small. On the other hand, it does not allow the ratio T_{max}/T_{min} to exceed the ratio W_{max}/W_{min} permitted by the radar. In the case of the *AN/APG-4*, W_{max}/W_{min} is about 5 to 1; in the *AN/APG-17* it might perhaps be extended to 10 to 1, because of the improved modulator used in that equipment. At the

cost of increased audio band and the use of at least one non-linear control element, counter range sensitivity h_R as well as radar range sensitivity k_R may be varied by the clock-setting servo. Equation (VIII.4) must still be fulfilled, but T_{max}/T_{min} may in this way be permitted to exceed W_{max}/W_{min}.

c. *Corrections.* If there is relative target motion transverse to the line of sight, as in the case of an aircraft flying so as to pass over rather than collide with its target, additional data beside R and S is required. If the aircraft is flying level and the target is seen at an angle α below the horizon (see Fig. V.-3), time to target is given by

$$(R/T)\cos^2\alpha - S = 0 \ . \tag{VIII.7}$$

This angle correction is the only effect of departure of the approach path from the line of sight to the target.

Using radar altitude A as additional data, the value of $\sin\alpha$ is A/R. An auxiliary computer determining α could be used to apply the proper correction to overall slant-range sensitivity. With such an addition, the basic tracker of Fig. VIII.-2 would solve equation (VIII.7) and give an exact solution for time to elapse until the aircraft passes directly over the target. This turns out to be an unduly complex way to solve the problem, however.

It was shown in section 3a of Chapter V. that a very close approximation to the exact relation between slant range and slant speed in level flight, for any given time to target, results from applying two simple corrections to equation (VIII.2). The corrected equation, (V.12), is

$$R/T' - (S - S_0) = 0 \ , \tag{VIII.8}$$

where the time factor T' differs from the approximate value for true time to target T only by a correction factor slightly smaller than unity, and where a fairly small negative speed intercept S_0 is introduced. These changes can be introduced into the radar tracker quite simply. Referring again to the diagram of the sweep-control circuit, Fig. III.-12, resistor r_2 can be fixed and output to the wave-shaping circuit can be tapped down on it by a small variable amount, to provide the appropriate time-

correction factor without disturbing the main reciprocal-time sweep-width control obtained through linear rheostat r_1. The servo motor can be made to hold total counter-output voltage to a value e_1 slightly greater than the bias e_0 applied to the counter load, thus providing for the required intercept S_0.

As shown in Chapter V., the time-correction factor T'/T and the speed intercept are both functions of only the single variable A/T. The functions required contain as parameters both upper and lower limits of the range of horizontal closing speeds within which time to target is to be approximated. After choice of the fixed limiting speeds to be used, the only additional datum required is the altitude A of level flight. The necessary correction functions, shown graphically in Fig. V.-5, are given by equations (V.14) and (V.22), using (V.9).

Fig. VIII.-3 shows in block form an arrangement providing slant-corrected time tracking. The clock-setting time-to-target servo is connected to the slant-range and slant-speed radar only as a sweep-width controller, just as it is in the basic tracker. Its time-to-target output shaft carries, however, an additional linear potentiometer that is fed from the regulated power supply of a radar altimeter and provides an output voltage proportional to T. This time voltage is applied in turn to a linear follow-up potentiometer driven by an auxiliary servo. The auxiliary servo operates to make its follow-up output voltage just balance the limit-counter output of the altimeter. Since the follow-up potentiometer with applied voltage proportional to T is driven to a position producing an output proportional to A, the servo follow-up shaft must evidently seek an angular position proportional to A/T. The auxiliary servo is thus able to apply to the slant-range radar the proper modulation-sweep and counter-bias corrections to make the main time-tracking servo read correctly.

Potentiometers with non-linear resistance-rotation characteristics of special shape are necessary to introduce the corrections as functions of A/T. Since these corrections are small, however, great accuracy of shape of characteristics is not necessary. Smallness of corrections has the further useful effect of promoting overall stability

of the two-servo system of Fig. VIII.-3. Since the value of A/T has only a slight effect on the balance point of the main T servo, there is no complicated interaction between

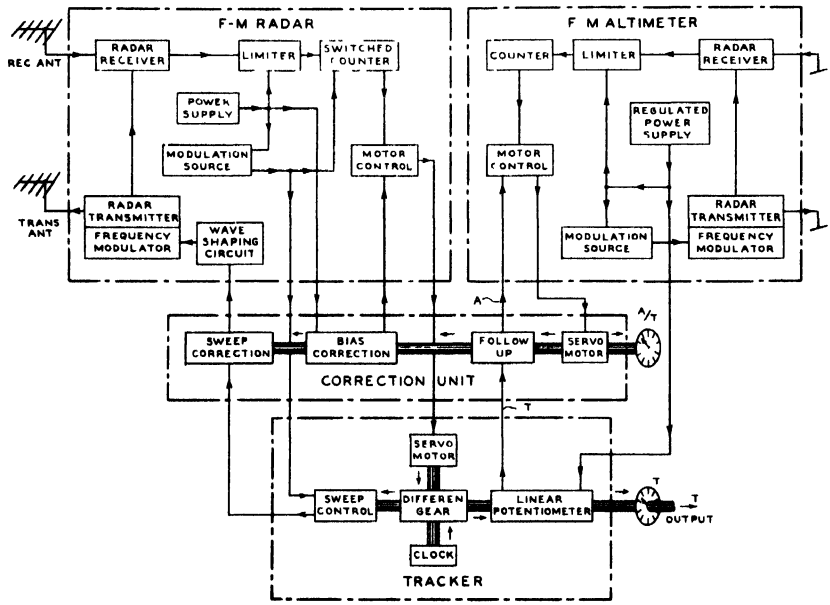

Fig. VIII.-3. Block diagram of time tracker with altitude correction.

the two servos and it is only required that each of them individually be made well damped and stable.

Restriction of time tracking to level-flight approaches is not necessary. Section 4a of Chapter V. has shown that exactly the form (VIII.8) of range-speed approximation remains valid even in the presence of a vertical component of aircraft speed. The additional datum of vertical speed V becomes necessary in that case, however, and the time-correction factor T'/T and speed intercept S_0 each depend on both A/T and V, as indicated by equations (V.32) and (V.35). The slant-corrected tracker of Fig. VIII.-3 requires only to be altered to provide such two-variable control of its corrections in order to remain useful for gliding or climbing straight-line approaches.

No correction for radar time lag is necessary. If the tracker is delivering the correct value of time to target

at the instant that a particular range datum is accepted, the servo will be so informed after a circuit time lag τ_s and will then take no corrective action. Since the clock will keep the T output correct once it has become so, the no-correction order will continue to be given to the servo. There will be delay in correcting an inaccurate value of T, but the correction applied after lapse of the circuit lag τ_s will be the proper one. To avoid over-correction and hunting of the T servo because of the time lag, however, a simple derivative-damping circuit must be applied. This circuit was omitted from Fig. VIII.-3 for simplicity; it may act to displace the balance point of the motor-control amplifier in proportion to the speed of the servo motor. Time lag in operation of the A/T servo will necessitate a very simple special correction. This is made by feeding the A/T follow-up potentiometer with a voltage proportional not to current time to target but to a time reduced by the lag to be corrected. When the A/T servo reaches the corresponding setting, after the lapse of its time lag, that setting will be correct for the time to target then current.

Residual range R_r in r-f transmission lines necessitates a special correction, as in other applications. This may be accomplished by changing the negative speed intercept, for which bias correction is made, from S_0 to $S_0 - R_r/T'$. Since the new correction depends upon T separately as well as upon A/T, the bias-control circuit becomes somewhat more complicated than that shown in the figure. The point to which time is determined may be shifted somewhat in range by correcting not merely for R_r but for $R_r - R_1$, where R_1 is the distance beyond the target of the point for which time of arrival is desired.

d. ***Automatic Controls.*** Sequence of operation is a very important factor in determining the success of an automatic tracker. Many possible types of operation can under some particular sets of conditions lead to failure to pick up a target, or at least to improper tracking. Complicated manual control procedures requiring considerable use of operator judgment are of course impracticable.

There is little advantage in using a sloping audio-amplifier gain-frequency characteristic, since variation

of range and consequently of signal strength takes place at constant beat frequency while tracking. The amplifier should therefore have a moderately flat band-pass characteristic, with wide enough pass band to accommodate only the minimum required range of closing speeds. This will give good signal/noise ratio when the target is being tracked, but the target will only be found at all if it happens to return an adequate signal while at those particular ranges that give frequencies in the amplifier pass band with the particular modulation sweep in use.

An acceptable operating sequence will be of the following sort. Upon first activating the system, manually or otherwise, a favorable condition for initial location of a target should be set up immediately. When a good signal is first received for a significant time interval, both servo tracking and clock timing should be started. Loss of signal thereafter should temporarily disable the servo until good signal is again available, but should neither disturb the clock nor reinstate the initial target-finding condition. At a time T_{min} before the target is reached, servo correction should be stopped and the remainder of the approach timed by clock alone. When the approach is terminated, either by arrival over the target or by prior operator control, the target-finding condition should be restored quickly and automatically. A new operating cycle on the next target found may then either be allowed to take place automatically, or if preferred a manual recycling operation may be required to permit subsequent tracking.

A simple target-finding condition is that of minimum modulation sweep. On a large target, signal in the amplifier pass band will then be picked up at long range and will initiate tracking when a time interval T_{max} has still to pass before the target is reached. A weak target will not provide adequate signal until the range is short, under which condition the beat frequency at minimum sweep may already be too low to pass the audio amplifier. Weak targets may therefore never be picked up at all if the sweep is held at minimum until a signal is found.

Targets will be found more reliably if the radar is made to search in range until a signal is picked up.

Search may be accomplished by causing the clock-setting servo to vary the modulation sweep steadily from maximum to minimum, return rapidly to maximum sweep, and then repeat the process. This will continue from the time the tracker is turned on until a radar beat strong enough to actuate a target-indicator relay is first brought into the amplifier pass band. Operation of the relay should then permanently stop the search action, start the clock, and cause the tracking servo to set the clock. Turning off the tracker control switch should recycle the system, so that search may again be started by again turning on this control switch. Rate of change of sweep width in search should be so chosen that on the one hand loss of time in finding the target will not be excessive, while on the other hand the interval during which the radar beat frequency remains within the amplifier pass band will not be too small for reliable relay operation.

If the tracker is to be used for bomb release, control circuits to cause release must be added. Release may be controlled on the basis of time of fall, which is related to altitude in accordance with equation (V.3) for the case of level flight. Time of fall T_f and the parameter A/T for level flight are therefore related in value by the equation

$$(2/g)(A/T) = T_f^2/T \ . \qquad \text{(VIII.9)}$$

The angular position of the A/T shaft, using the fixed scale factor $2/g$, increases according to (VIII.9) as time to target decreases during an approach.

When time to target becomes equal to time of fall, the angular positions of the two servo shafts coincide and the bomb should be released. Release fully compensated for altitude and closing speed may therefore be obtained automatically from a time tracker with practically no extra equipment. It is merely necessary to arrange for a contact to be closed when the T and A/T shafts simultaneously reach identical angular positions. This contact may be produced directly by mechanical means, or indirectly by comparison in a differential relay of two voltages controlled respectively by T and $(2/g)(A/T)$.

e. ***Experiments.*** An *AN/APG-4* Sniffer was modified to form the basic tracker of Fig. VIII.-2. Extensive bench

tests and limited flight tests of this unit yielded interesting information. Because of low priority of the tracker at the time it was tested, the *A/T* servo to correct for slant of the line of sight was never added. Automatic sequence control was tried, however, in a simple form setting up minimum sweep upon activation and maintaining that condition until a target signal strong enough to actuate a control relay is received. The experimental unit was designed to work from 12½ to 2½ seconds to target, with servo action automatically stopped at the latter value and tracking done by clock alone thereafter.

Laboratory tests made it clear that avoidance of all stray loading effects in the modulation-control circuit is essential to prevent serious departure from linearity of the scale of time, or reciprocal sweep width, of the tracker. In checking linearity and calibration, especial care is necessary to avoid obscure fixed-error effects.

Flight tests were made in which time tracking was checked both by reading the time-to-target scale when passing over a target and by automatic release of bombs in flight at a predetermined altitude. Both manual and automatic control of operating sequence were tried. Manual control would of course not be of practical value, but is a convenient way of making special tests.

Errors were observed which corresponded in a general way with those expected from lack of slant correction. Some tests were therefore made in which the servo was shut off manually at a considerable range, where the slant correction was small, and the remainder of the approach tracked by clock alone. This was of course a condition tending to produce large instrumental errors, but nevertheless resulted in several strikingly good approaches.

Only timed runs were possible on good targets, and a number of these yielded timing accurate within the error of observation, which amounted to a few tenths of a second. Automatic initiation of tracking from minimum-sweep modulation setting was found entirely successful on targets of adequate size. On the weak target used for bombing, however, automatic starting from minimum sweep (maximum range) sometimes failed for lack of early signal; automatic starting from reduced range would have been satisfactory,

as would automatic starting from a searching condition.

Operation of the anticipated type was observed in flight on a number of runs, with radar and servo acting immediately to set the clock upon first reception of strong signal, and the clock thereafter tracking correctly for a number of seconds with the setting servo idle. Fading of the signal during clock tracking then produced no disturbance.

Increased useful bombing accuracy may be expected with a satisfactory time tracker, because the servo-actuating signal threshold may be set relatively high without losing operation altogether. Radar operation will then take place only at moments when the signal is exceptionally clean and so provides reliable data. Another source of improvement of accuracy by tracking is the fact that many targets show strong fading in just the range region of low-altitude bomb release; the tracker effectively eliminates disturbance of release by fading, provided an accurate clock setting has once been established. No further development of trackers was undertaken for lack of specific applications requiring them.

6. MEASUREMENT OF VERTICAL SPEED

a. *Principles of Simple Methods.* It is desirable to be able to derive from a radar altimeter indications of the vertical speed of aircraft. Such indications might prove directly useful to the pilot, because the air-viscosity rate-of-climb indicators which he must use at present are extremely sluggish in action. Vertical-speed data is also necessary if time to reach target or time of fall of bombs is to be determined for an aircraft that is not held in level flight.

Use of Doppler frequency shift of course suggests itself as a means of measuring vertical speed. In normal aircraft operation, however, vertical speed may be either upward or downward; zero or almost zero values are of common occurrence and particular importance. This means that very small differences between large upsweep and downsweep beat frequencies would have to be measured in Doppler speed determination with a frequency-modulated radar signal.

In the **AN/APN-1*, for example, the difference between upsweep and downsweep beat frequencies for a climb of 100 feet per minute would be only 2.7 cycles per second. The average of upsweep and downsweep frequencies would be 7300 cycles per second at an actual altitude of 350 feet (with additional residual altitude 34 feet). The balance requirement to be met by a switched speed counter, for even roughly correct speed indication under such conditions, is obviously of an impracticable order. The Doppler-shift method therefore seems unpromising for rate-of-climb measurement with f-m radar altimeters of present design.

A more promising but less direct line of attack is development of vertical-speed information by differentiating altitude with respect to time. Since f-m radar altimeters are capable of developing by simple means fairly well smoothed output voltages varying rapidly with altitude, such methods seem entirely practical. The non-linear indicator circuit of the **AN/APN-1* altimeter develops, on the low scale, an average voltage variation of 0.13 volts per foot at the cathode of its cathode follower. At 100 feet per minute of climb, this voltage varies 0.21 volt per second. With a smoothly acting servo to balance the null counter of the altimeter limit circuit, the voltage at the arm of the follow-up potentiometer in that circuit may vary 0.56 volt per second for the same rate of climb.

There are various ways in which electrical differentiation can be applied to these voltage variations to produce indications of vertical speed. The simplest is of course the connection of a capacitor and galvanometer in series across the variable voltage. As voltage e varies, a charging current $C^{de}/_{dt}$ will flow into the capacitor C and deflect the series galvanometer. At 100 feet per minute, variation of null-counter balancing voltage at 0.56 volts per second will produce a current of 1.68 microamperes if a 3-microfarad capacitor is used. For a device giving full-scale deflection at vertical speeds of ±6000 feet per minute, a standard range for military rate-of-climb indicators, the galvanometer required will then be a 100-0-100 microammeter; this is a moderately rugged instrument.

Very rapid indication is obtained with this simple circuit; the time lag involved is that of the altimeter

and servo in developing a voltage proportional to altitude, plus the mechanical time constant of the galvanometer itself. Null-counter balancing voltage being strictly linear with altitude, the circuit gives a linear variation of galvanometer current with vertical speed and the speed-scale factor is independent of altitude. If the follower-cathode voltage of the *AN/APN-1* indicator circuit is used as data source instead of the limit-circuit balancing voltage, a non-linear d-c amplifier must be introduced to cancel the non-linearity of the indicator counter in order to obtain linear vertical-speed indication independently of altitude. Direct use of the non-linear follower-cathode voltage would result in speed indications varying linearly with vertical speed but weighted according to altitude, a condition useful for special purposes.

If a really rugged instrument must be used for speed indication, or if a voltage or shaft rotation proportional to speed is required instead of a meter indication, the differentiating circuit must be modified. A modification that can be applied in any one of several ways is that of Fig. VIII.-4, in which the current i charging capacitor C passes through a load resistor r. The most commonly used differentiating circuit is obtained by short circuiting the e_2 terminals.

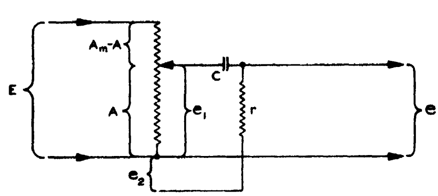

Fig. VIII.-4. Differentiating circuit.

Solution of the differential equation of the circuit of Fig. VIII.-4 with its e_2 terminals short circuited shows some typical properties. If input voltage e_1 has been constant but suddenly begins to increase linearly with time, the output voltage e which has been zero will simultaneously begin to increase. e will approach exponentially a final constant value $rC\,de_1/dt$, with a time constant rC in the exponent. If e_1 has been constant and begins suddenly to increase at a uniformly accelerated

rate, e will again begin simultaneously to increase from zero. In this case, however, the ideal value $rC\,{}^{de_1}\!/_{dt}$ for e will increase linearly with time, and the steady-state value that e approaches exponentially with time constant rC will be less than this ideal value by the constant defect $r^2C^2\,{}^{d^2e_1}\!/_{dt^2}$.

Taking into account the linear variation of e_1 with altitude A, indicated by the potentiometer input shown, the sensitivity of the circuit to vertical speed V is easily found. This sensitivity e/V is simply ErC/A_m, as is evident from the relations of ${}^{de_1}\!/_{dt}$ to V and to e. In the uniformly accelerated case, this value of speed sensitivity still holds, and the output-voltage defect found corresponds simply to a fixed time lag of output voltage behind input speed, equal in amount to the time constant rC. The value of rC thus controls three things: sensitivity to speed, rate of decay of disturbing transient circuit conditions, and time lag in following a uniformly accelerated speed. With e_2 terminals shorted, a large value of rC to give high sensitivity necessitates slow decay of transients and a large lag in following accelerated motion.

b. ***Feed-Back Methods.*** A direct-current polarity-reversing amplifier with gain G may be introduced between the e terminals and the e_2 terminals of Fig. VIII.-4, to make

$$e_2 = -Ge \tag{VIII.10}$$

with no significant time lag. Solution of the circuit equation in this case shows that e behaves in just the same way as with e_2 shorted, except that the effective time constant is reduced from rC to $rC/(1+G)$. The voltage e_2 varies with time just as does e but is G times greater than e; the sensitivity e_2/V in the circuit with amplifier is therefore $(ErC/A_m)G/(1+G)$, which is just $G/(1+G)$ times the sensitivity e/V of the simple circuit with e_2 shorted. That is, by introducing an amplifier of high gain and using e_2 as speed-sensitive output, high sensitivity may be obtained through use of a large value of rC, while at the same time the effective transient-decay time constant and following lag $rC/(1+G)$ may be kept small to give very rapid response.

A somewhat different type of feed back results if e_2 is supplied by a potentiometer driven by a servo which is controlled by e but has strong first-derivative damping. This gives a viscous behavior, so that

$$de_2/dt = e/\tau \qquad \text{(VIII.11)}$$

where the time constant τ is the primary characteristic of the servo system. The differential equation of this system is of the second order and so has a more complicated solution than the previous cases.

The same two cases, that of sudden appearance of a constant vertical speed and that of sudden application of constant vertical acceleration, again serve to give a good idea of the properties of the system. Sensitivity e_2/V again proves to be ErC/Am. Output voltage e_2 again follows a uniformly varying vertical speed V with a uniform time lag, after a steady state is reached, but this time lag has become independent of rC and is the time constant τ of the damped servo. Decay of transients depends on both τ and rC, but unless the servo is over-damped and sluggish, transient decay is practically exponential with time constant $2rC$. If τ is less than $4rC$, the decaying transient is oscillatory.

It is necessary in all cases that variation of e_1 shall be reasonably smooth. Ripple or unsteadiness of e_1 will produce large alternating components at the output of a sensitive differentiator, which may easily overload some part of the system and prevent proper operation unless suitable precautions are taken. So long as overloading is avoided and the residual ripple or noise on the vertical-speed output voltage e or e_2 is not harmful to the utilization of that voltage, ripple or noise on e_1 may be tolerated. Use of a high-quality capacitor is always important. Leakage conductance or dielectric after-effects in the differentiating capacitor will of course invalidate the above discussion and seriously impair the operation of any differentiator.

Given a sufficiently smoothly operating altitude servo, a tachometer driven mechanically by that servo may be used instead of an electrical differentiator. An electrical generating tachometer so driven, for example, will produce

an output voltage directly proportional to vertical speed.

c. *Experimental Results.* Low priorities assigned to derivation and utilization of vertical-speed data severely limited the experimental work in this field. Considerable laboratory experimentation was nevertheless accomplished and a very few flight tests were made.

In one case the non-linear indicator-circuit output of the **AN/APN-1* was used to give a vertical-speed signal weighted according to altitude. This was for the purpose of making first-order corrections in bomb release, and will be described further in that connection in the next section. Operation was required in conjunction with a limit-circuit servo of the start-stop type (*SA-28/APG*) having a marked idle region. Laboratory and flight test of this vertical-speed circuit showed it to be operative in principle but badly disturbed in practice by transients caused by starting and stopping of the servo used. This difficulty, for which no complete remedy was found in the time available, prevented completion of the development.

In exploring the possibilities of altitude and vertical-speed indication, a servo with operation smoothed by small forced oscillations, as described in section 5c of Chapter IV., was applied to balance the limit circuit of an **AN/APN-1*. This made available a relatively rapidly and smoothly varying follow-up voltage linearly proportional to altitude, with only moderate ripple. A simple rC differentiator (Fig. VIII.-4 with e_2 shorted) was connected to this voltage, and the differentiator output (e of the figure) was applied to the grid of a pentode current amplifier with a rugged milliammeter in its cathode circuit.

By careful design, it was possible to obtain a total cathode-current variation of 5 milliamperes (full scale of the indicating meter) for a vertical-speed range of ±2000 feet per minute. This sensitivity, at a time constant of 1.5 seconds, permitted so little cathode-circuit degeneration in the current amplifier that considerable care was necessary to avoid undue disturbance by supply-voltage variations. Some electrical filtering was necessary to avoid vibration of the indicating meter at the very low servo-oscillation frequency.

This instrument operated well in flight, but in the limited maneuvers possible with the *SNB* aircraft used little significant difference was noted between its indications and those of the normal 5-second air-viscosity instrument. Rapid response of the radar rate-of-climb indicator was easily observed in the case of aerodynamic "bumps" in the flight of the aircraft, to which the air-viscosity instrument did not respond at all. Since no application for this instrument was evident, no further work was done on it.

Considerable preliminary laboratory testing was also done on servo-feed-back differentiators with viscous-type damping, in an effort to develop a source of vertical-speed data in the form of shaft rotation for use in fire-control computers. This work showed promise, but could not be continued beyond the early stages.

Mention may be made here of the importance and difficulty of producing a voltage varying truly smoothly at a known and controllable rate. Such a voltage is quite necessary to simulate the d-c altitude signal produced by vertical motion of an aircraft, if laboratory testing and calibration of vertical-speed measuring equipment is to be at all effective. No truly satisfactory solution to the problem of producing a test voltage was found in the course of the work on f-m radar determination of vertical speed. Cyclic variation of simulated altitude is necessary for reasons of practical convenience, and must take place without exciting marked transients in the system under test. A carefully constructed motor-driven potentiometer of special design, giving alternating periods of constant simulated speed and constant acceleration, was the most useful test source developed.

7. Bombing in Vertical Maneuvers

a. *Purpose.* Low-altitude bombing from level flight has two disadvantages. It requires very careful flying, by instruments if conditions of poor visibility make radar truly useful, and it makes the path of the bombing aircraft simply predictable and so makes that craft a good target for gun fire. Dive bombing reduces target motion across the line of sight and makes the missile trajectory

straighter, both factors leading to improved accuracy and increased danger to the attacker. Toss bombing permits release at relatively long ranges, increasing the safety of the attacker but decreasing accuracy.

To remove the requirement for careful level flying, it is merely necessary to arrange that small vertical speeds shall not introduce serious range error. To achieve maximum safety of aircraft, it is necessary to go further and to give the pilot as much freedom as possible to maneuver as he sees fit in the vertical plane of approach. Work was started leading toward both these objectives, but had to be dropped before completion in both cases.

b. *First Approximation.* Bombing errors caused by small departures from level flight can be largely corrected by simple approximate methods. The kinematic basis of one such method has been developed in section 4d of Chapter V. Correction by this method requires only that the bombing computer of a normal Sniffer be compensated for a modified altitude rather than for that at which the bombing craft is actually flying. Equation (V.44a), using the value for V_0 given by (V.43), indicates that the altitude compensated should exceed the actual altitude by a definite fraction determined by the single variable $V/\sqrt{A}$, where V is rate of climb and A is altitude.

The variable $V/\sqrt{A}$ is simply twice the time rate of change of $\sqrt{A}$, as may be seen by differentiating the latter. By a fortunate chance, the cathode voltage of the cathode follower in the non-linear indicator circuit of the **AN/APN-1* altimeter is found to be very closely proportional to the square root of altitude, at least over the important altitude range of 100 to 400 feet. The time derivative of this voltage therefore has just the right form to be used in modifying altitudes within that range to compensate bomb release for small amounts of vertical speed. Modification may be accomplished through control of the voltage supplied to the follow-up potentiometer of the altitude servo in a normal *SA-28/APG* compensation unit. By subtracting from the fixed supply voltage a variable voltage proportional to time rate of change of altitude-indicator current, and applying the difference to the follow-up potentiometer, the servo may be made to seek a

position representing the modified altitude A' of equation (V.45a) rather than the actual altitude A.

The approximation (V.45a) is quite good up to a vertical speed of $\pm\frac{1}{5}V_0$ (V_0 is the final vertical speed attained by a bomb released in level flight), which is 16 feet per second or almost 1000 feet per minute for an altitude of 100 feet, or 2000 feet per minute at 400 feet. At an altitude of 225 feet (V_0 of 120 feet per second) and a rate of climb of ±10 feet per second, the correcting increment required for a fixed supply of 150 volts is ±21.2 volts. The rate of change of indicator-amplifier cathode voltage in the *AN/APN-1* altimeter under these conditions is ±1.26 volts per second. Using a simple differentiating circuit with ½-second time constant, the control voltage available is ±0.63 volt. A phase-reversing direct-current amplifier with voltage gain of 34 is therefore required to provide the modifying component of follow-up supply voltage.

An experimental accessory unit built to correct in this way for vertical speed used the circuit of Fig. VIII.-5.

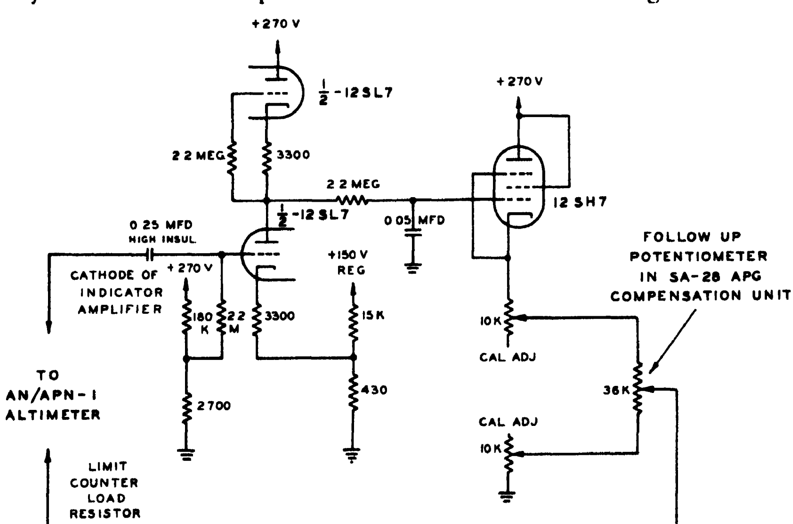

Fig. VIII.-5. Circuit of accessory unit to compensate bomb release for limited vertical speed.

The d-c amplifier, using a constant-current type of plate load and having its grid biased from the plate supply, is

not affected by plate-supply variations. During limited tests of this basically sound unit, satisfactory operation was prevented by transient disturbances of the regulated supply voltage caused by servo operation. It should be noted that the two-tube correcting unit shown can be connected between an altimeter normal except for such connection and an entirely normal *SA-28/APG* unit, for use with entirely normal *AN/APG-4* bombing equipment. This is another instance of the ease with which f-m radar equipment lends itself to automatic control or compensation.

Another approximate method of correcting for vertical speed is the use of a suitable increment to time of fall. Using equation (V.43), equation (V.36) indicates that, so long as V/V_0 is sufficiently small, time of fall may be regarded as the sum of a portion V_0/g depending only on altitude and a portion V/g depending only on vertical speed. If the time lag τ_s between occurrence of the release relation of slant range R to slant speed S and release of a bomb by the Sniffer can be controlled, correction by time-of-fall increment becomes possible. Such a result follows if Sniffer operation does not directly release a bomb but instead initiates a delay cycle of controllable duration, at the end of which the bomb is released. The additional delay must simply be made proportional to vertical speed in order to provide the desired correction.

Vertical-speed information for use in controlling time of fall directly must of course be derived from a linear altitude indicator. Time lag may be controlled by using the vertical-speed data to vary the threshold voltage at which the charging of a capacitor, initiated by Sniffer relay operation, actuates a bomb-release relay. This method of correction is also rather simple if a well-smoothed source of linear altitude data is available; the Sniffer computation must of course be altered to provide sufficiently early relay closing. It will not work without modification of altimeter and *SA-28/APG* unit, however.

Both the above approximate methods of correcting for vertical speed neglect the variation of slant speed with vertical speed and take account only of the effect on time of fall. The additional correction for the effect of vertical speed on slant speed is usually decidedly smaller,

except for especially low slant speeds. When it is necessary to use this additional correction, it is most easily applied at a different point in the circuit. One way of doing this is to increase the speed intercept S_0 of the bombing approximation by a factor $1+2V/V_0$; this is of course to be done by operating upon the counter-bias circuits of the *RT-27/APG-4* or *SA-28/APG*.

c. *Second Approximation.* High enough vertical speeds to permit evasive action may be used, without in principle sacrificing the accuracy obtained in level flight, by applying the results of sections 4a, 4b, and 4c of Chapter V., with proper allowance for the windage corrections of section 6 of that chapter. This calls for adjusting both modulation sweep and counter bias of the Sniffer in accordance with both altitude *A* and vertical speed *V*, making due allowance for the effect of vertical acceleration *a* during the various time lags encountered. Two methods of making such adjustment will be described in some detail, as the only cases in which application of other than the most elementary computers to f-m radar has been studied. Provision for minor corrections adds considerable difficulty to the design and mechanism of such computers, and will be given proportionate attention here.

An experimental Sniffer was built to use a two-variable compensation unit in flight trials of automatic bombing during vertical maneuvers.[1] A normal *RT-98/APG-17* radar transmitter - receiver (see Figs. VI.-21 and VI.-22), operating at 1500 megacycles, was to be used in conjunction with a special power and computing unit. The computer unit as built uses substantially the circuits of the corresponding portions of the *AN/APG-4*, but is arranged to include mechanically a two-variable compensator with its two driving servos. Smooth altitude control was to be derived from the limit circuit of an **AN/APN-1* altimeter by a servo in forced vibration, like that used in the *AN/APG-17A* equipment. Vertical-speed control was to be derived from altitude-servo follow-up voltage by a servo-balanced differentiating circuit. The release compensator was to develop the sweep-control and bias-control parameters T' and *So* of the bombing approximation; the same T' parameter was to compensate both the modulation sweep and

the scale of a manual range-lead adjustment.

Any reasonably behaved function of two independent variables may be developed by use of a three-dimensional cam. Such a cam may be a properly shaped solid, rotated about a fixed axis in proportion to one independent variable and translated along that axis in proportion to the other variable; the desired function is then developed by the varying radius of the cam on a fixed pick-off line. Construction of an original three-dimensional cam is a tedious job of high-precision machine and hand work, but copies may be made automatically. The mechanism necessary for using such a cam to develop an arbitrary function is quite simple.

If a function to be developed has a large component of variation that is monotonic with one or both independent variables, that function may be decomposed into linearly and non-linearly varying components. Precision may then be improved or cam size reduced by using the cam to develop only the non-linear variation and producing the linear components by simpler mechanical couplings, arranged to give suitable motion ratios.

Two three-dimensional cams were expected to be required in the experimental compensation unit, to provide the necessary T' and S_0 outputs in response to the A and V input data. The sweep-control parameter T' of course depends on time of fall T_f [see equation (V.36)], on Sniffer time lag τ_s, and on a correction factor $T'/(T_f+\tau_s)$ for obliquity of flight path with respect to radar line of sight. Obliquity corrections $T'/(T_f+\tau_s)$ and S_0 [see equations (V.38) and (V.40)] are quite small and need not be developed with extreme accuracy. Time of fall, on the other hand, is the primary release-controlling datum and must be determined with the highest practicable accuracy. This means that in the experimental unit the cam-input data has to be effectively the altitude and vertical speed of the aircraft at the instant of bomb release, in order to determine correct time of fall. Actual data collected, however, is that for appreciably earlier instants, because of altimeter, differentiator and Sniffer time lags. Means of accomplishing to a sufficient degree of approximation the data prediction required, and of allowing for bomb trail due to wind resistance, will be discussed later.

Fig. VIII.-6 is a perspective view of the experimental compensation unit as actually built. It proved advantageous in the T' system to separate out linear components of variation with both A and V. These are added by differential gearing and applied to the arms of ganged rheostats controlling both sweep width, or radar-range scale, and range-lead scale. The non-linear T' component from the three-dimensional cam is added to the sum of the linear components by causing the cam follower to move the frames and windings of these rheostats.

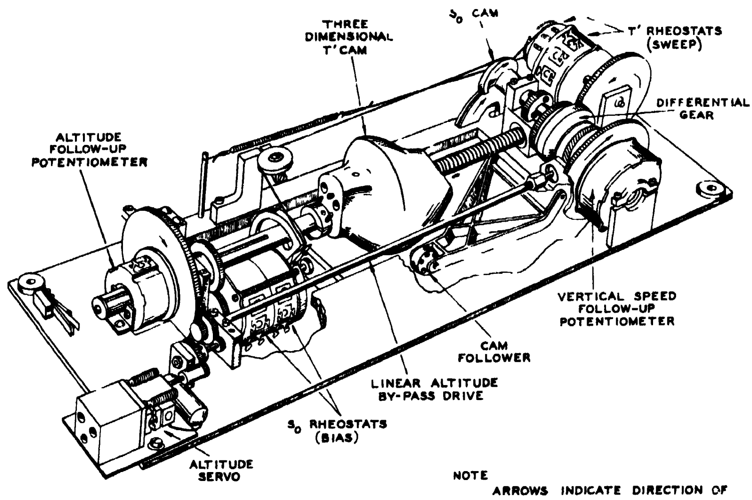

Fig. VIII.-6. Pictorial view of experimental two-variable release-compensation unit.

After separating out the linear dependence of S_0 on altitude, and driving in accordance with it the arms of ganged bias rheostats, the remaining non-linear altitude dependence of S_0 proved insufficient to justify use of a second three-dimensional cam. Nor did required accuracy justify separation of the large linear component of vertical-speed dependence of S_0, so a simple flat cam is used to rotate the frames and windings of the bias rheostats in accordance with the entire variation of S_0 with V.

The unit shown is obviously not a finished design. It is wasteful of space because allowing room for two three-dimensional cams where only one is used, since the complete two-variable bias-compensation portion of the unit proved

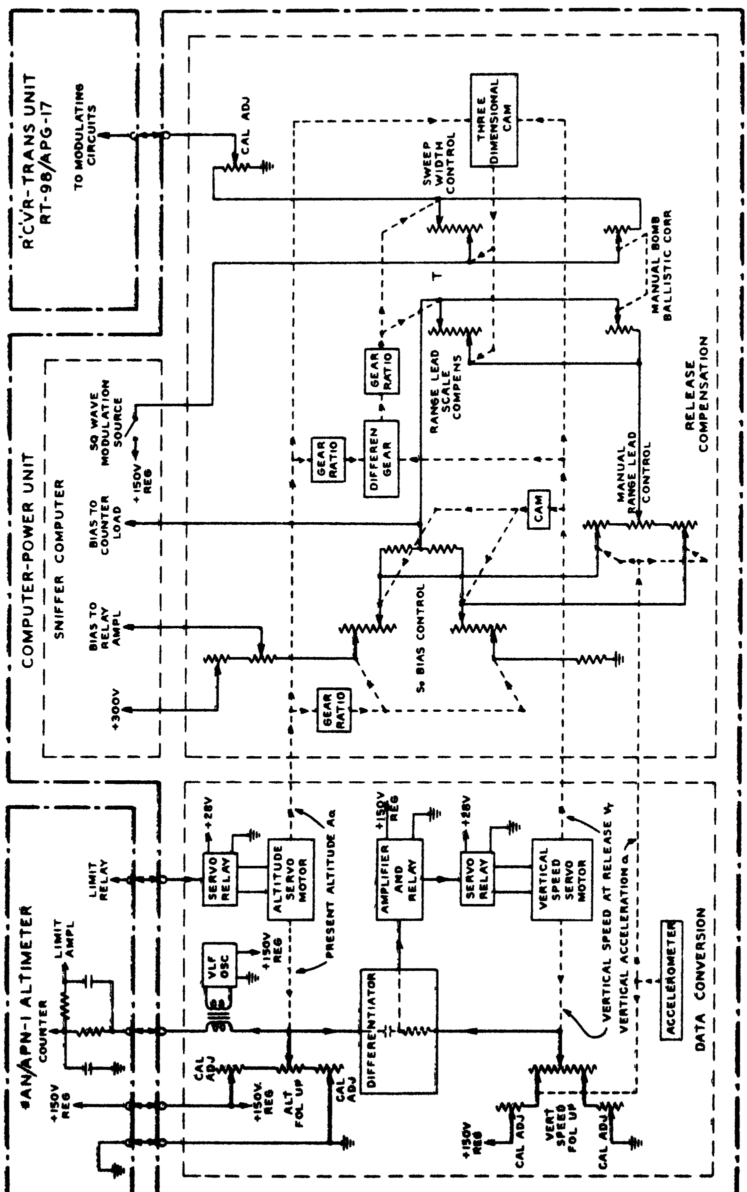

Fig. VIII.-7. Block schematic diagram of compensation unit for bomb release in vertical maneuvers.

unexpectedly simple in its mechanical arrangements. It nevertheless serves to illustrate a number of mechanical problems typical of more advanced fire-control devices than the original Sniffer, and to indicate simple constructional features for their solution. Fig. VIII.-7 is a block-schematic diagram of the two-variable compensation unit. Gear ratios are indicated where they are required to introduce scale factors. Cams, beside providing desired curve forms, act like gear ratios to permit establishment of scale factors where necessary.

The solid cam is mounted on a slidable shaft, machined with a spline on one end and a screw thread on the other end. Rotational drive from the altitude servo is applied to the spline by a keyed gear. Lengthwise drive from the vertical-speed servo (invisible below the bed plate) is applied by a worm gear meshing with the screw thread on the cam shaft. Cam output is taken from a spherical follower swinging in an arc in a radial plane through the shaft axis. To avoid close tolerances on centering of shafts, precise rotation of less than a full revolution is transferred from the solid-cam follower to the cases of the T' rheostats by cable and drums rather than by gearing. The flat cam on the vertical-speed shaft likewise drives the cases of the S_0 rheostats through cable and drum. Linear altitude drive is transferred through a shaft and gear ratio to the vertical-speed mechanism, where it is added to linear vertical-speed drive in a differential gear and applied to the shaft of the T' rheostats through another gear ratio. Linear altitude drive is likewise applied to the shaft of the S_0 rheostats through a gear ratio. The altitude servo drives an altitude follow-up potentiometer directly through suitable reduction gearing, just as the hidden vertical-speed servo does the shaft of the vertical-speed follow-up potentiometer. The case of the vertical-speed follow-up potentiometer was to have been driven by an accelerometer (not shown in Fig. VIII.-6) for time-lag correction.

Use of a screw thread as a rack for moving a solid cam lengthwise is convenient but causes slight undesired lengthwise motion as the cam rotates. This does no harm if the cam is cut with each contour of constant vertical

speed not lying in a plane normal to the cam-shaft axis, but rather advancing lengthwise with the same pitch as the driving screw when the cam turns. Altitude indicated by rotation of the cam shaft is not that at release, but rather that measured prior to release by an interval equal to the total time lag in altimeter, altitude servo, Sniffer and release mechanism. This does no harm if the cam is cut with each contour of constant altitude not lying in a plane through the cam-shaft axis, but rather in a helix twisted at a rate proportional to the total time lag. Thus cut, the cam surface in effect rotates under the follower for a constant input altitude, even though the cam shaft does not, as vertical speed sets the lengthwise position of the cam. Such phantom rotation is used in the experimental compensator to apply proper correction for the action of vertical speed in changing altitude during the total lag time.

Analytical correction of a solid-cam surface for finite follower size would be very difficult. This is avoided by using a spherical follower, large enough not to indent the cam, and cutting the cam with a spherical cutter having exactly the radius of the follower. Using this method of cam development, radii on which the cam is to be cut are simply specified to the center of the cutter, and allowance for motion of the point of cam contact on the spherical follower surface occurrs automatically in the process of cutting the cam. Motion of the follower center along a circular arc rather than along a single radial line produces further complication. Corrections to cutter-center radius and to longitudinal position for arcuate follower motion are, however, easily made analytically when designing the cam.

Not all of the possible cam surface is used, since some combinations of permitted altitude and climb lead to excessive release ranges. Some of the otherwise unused surface is made cylindrical and provided with a "bench mark" to establish simultaneously definite, known values of altitude, vertical speed, and angular position of follower arm. The bench mark is essential as an aid to initial adjustment of the compensation unit.

Vertical acceleration acts primarily to alter vertical

speed of the aircraft during the interval between speed determination and release. Allowance for this was to be made in the experimental unit by using a vertical-component accelerometer of the mass-and-spring type to rotate the frame and winding of the vertical-speed follow-up potentiometer, through an angle proportional to the product of vertical acceleration and time lag. A similar correction could have been made to the altitude follow up for the $\frac{1}{2}a\tau^2$ altitude increment, but was not considered essential. Making S_0 and $T'/(T_f+\tau_s)$ corrections for the moment of release rather than for that of "sniffing" also produces slight errors dependent on acceleration. It was found graphically that a useful but very rough overall correction both for these errors and for altitude increment could be made by applying a small range-lead increment proportional to acceleration. Circuit provision for this additional range lead is indicated in Fig. VIII.-7.

Horizontal acceleration was neglected in planning the experimental equipment, and is to be avoided by care in flying. Vertical acceleration alone was to be measured, even in steep climbs or dives, by hanging the accelerometer mass as a plumb bob. The accelerometer was never completely constructed and is not shown in Fig. VIII.-6; it would have been mounted at the front of the bed plate, next to the vertical-speed follow-up potentiometer.

Bomb trail resulting from air resistance, neglected in earlier level-flight Snifler equipments, becomes significant at the longer times of fall encountered in toss bombing, especially at maximum flight speeds. Trail corrections do not become so great under vertical-maneuvering Sniffer conditions, however, that they must be made with great accuracy. Variation with speed of bomb trail in range is non linear, but a linear approximation over a limited speed range leads to no great errors.

The linear trail approximation may be characterized by its slope, T_t, and a speed intercept. Trail slope T_t is found empirically to depend very little, for any single time of fall, on altitude and vertical speed separately. It does of course depend upon ballistic coefficient of the bomb used, and this dependence on bomb type prevents full trail allowance from simply being made once and for

all in the design of the solid cam for T'.

It is not the speed intercept of the trail-approximating line itself that matters, but rather the change that inclusion of trail produces in the intercept S_0 of the main range-speed approximating line. This change in intercept depends on bomb characteristics as well as on both altitude and vertical speed. Its magnitude is small, however, so variation of intercept over a moderate range of bomb characteristics is also small. An intercept correction made once and for all for a bomb of average ballistic quality is therefore good enough for bombs having properties moderately close to that average.

Trail-approximation slope T_t (always negative) adds directly to the slope T' of the main range-speed approximation to determine a final operating value $T'+T_t$, and thereby the radar-modulation sweep, width required. Sweep control in proportion to $1/T'$ may be produced by application of modulating signal at constant voltage to a simple voltage-dividing series circuit, in which a servo-adjusted rheostat maintains total resistance proportional to T' and output is taken across a fixed resistor (see Fig. III.-12, but with r_2 fixed). It turns out with such a circuit that the trail-slope correction required to make sweep width proportional to $1/(T'+T_t)$ can be closely approximated for any single bomb by shunting a fixed resistor across the rheostat. The size of the fixed series resistor delivering output and that of the fixed shunt must both be properly chosen, with respect to the time/resistance scale of the variable resistor, if the best trail approximation is to be obtained in this way.

The method used in designing the experimental vertical-maneuvering Sniffer compensator to allow for trail involved several steps. First, series and shunt fixed-resistor values required in the sweep-control circuit to give best average trail-slope correction were determined, for the case of sweep-control rheostat rotation proportional to T' for vacuum fall [as found from equation (V.38)] and use of a bomb of medium ballistic coefficient. Using the circuit values so found with the vacuum-fall values of T' and with values of T_t obtained from computed trajectories, the required rotation angles for the sweep-control rheostat

were next determined exactly for all values of altitude and vertical speed at release.

Lengthwise profiles of cam radius versus vertical speed were thus determined for many fixed values of altitude. The same data could of course be regarded instead as a set of circumferential radius-versus-altitude profiles for many fixed values of vertical speed. Applying to these profiles the necessary corrections for arcuate follower motion, for prediction of altitude at release, and for screw-thread longitudinal drive, the coordinates of many points on the cam surface (using radii to center of follower) were finally tabulated for use in actual cutting of the cam.

The result of the above procedure is a solid cam giving with shunted rheostat optimum values of sweep width at every altitude and vertical speed for a bomb of medium ballistic coefficient, yet differing very little from the cam that would have been required for best vacuum-fall operation (ballistic coefficient infinite) with no shunt on the sweep-control rheostat. By varying the ganged manual bomb-trail correction rheostats which shunt the servo-driven sweep-control and range-lead scale rheostats of Fig. VIII.-7, range-speed slope can be made very nearly correct at all altitudes and vertical speeds for other bomb types having slightly different ballistic coefficients. Variation of speed intercept S_0 with altitude and vertical speed is also made correct for bombs of medium coefficient; the change of intercept called for by small changes of ballistics is neglected.

The problem of accurate release of bombs subject to air resistance from an aircraft maneuvering freely in a vertical plane is a complex one. A number of approximations had to be made in order to handle it in the simplified way just described. The design of the required solid cam was a tedious procedure requiring a number of corrections. Yet the end result, in consequence of the great flexibility of three-dimensional cams and of the ease with which f-m radar data may be put in forms useful for automatic control, is a release-computing system that is relatively simple both electrically and mechanically.

Operating ranges for which the two-variable computer

is designed are from 100 to 800 feet in altitude, ±150 feet per second or 9000 feet per minute in vertical speed, and from 1g downward (free fall) to 2g upward in vertical acceleration, for horizontal speeds of closing on target from 120 to 350 knots (200 to 600 feet per second). In addition, time-of-fall limits of 2.3 and 11.8 seconds are imposed. Flight maneuvers under downward acceleration exceeding 1g are permissible, as noted in section 4c of Chapter V., but actual release simply cannot occur during such a maneuver. Upward acceleration in excess of the computer-design limit, on the other hand, does not prevent release but does impair accuracy.

The vibrating altitude servo was found able to operate smoothly and accurately at the speeds required. There was indication, from laboratory servo tests and from flight test of the vertical speed indicator described in section 6c above, that a satisfactory vertical-speed servo operating from the altitude follow-up voltage is possible. Priority of other work unfortunately prevented the maneuvering Sniffer development from reaching the stage of complete system tests.

It seems highly probable that, given suitable vertical-speed data, the two-variable compensator would have operated as expected. That slant-range and slant-speed radar data in maneuvers would have been consistently good enough was never clearly established. Limited observations on 410-megacycle radar signals in dives, made in connection with studies of rocket sighting, and still more limited observations when climbing, were somewhat encouraging with regard to signal quality in moderate maneuvers. The weakness of signals from the existing 1500-megacycle equipment at long range, observed in level-flight rocket firing, was not particularly encouraging. Measures to improve the radar performance would probably have been necessary to make an improved computer truly useful. Pitch stabilization of antennas might have proved necessary also.

d. *An Alternative Approximation.* Design of a solid cam with all necessary corrections is a very lengthy and tedious procedure, as is the actual construction of a prototype cam to adequately close tolerances. Even minor changes in operating conditions may necessitate complete

cam revision. Much thought was therefore given to possible ways of building two-variable release compensators requiring no such difficult element. The schemes devised were guided by the mathematical forms expressing the quantities T' and S_0 that were to be determined. All of these schemes are characterized by use throughout of fairly easily designed and constructed elements, but all of them require a considerable number of such elements to produce a complete compensating unit.

Time of fall is the most important single variable in Sniffer compensation, so the alternative schemes studied were based on maximum explicit use of that variable. Plotting range-speed slope-correction factor $T'/(T_f+\tau_s)$ and speed intercept S_0, as given by equations (V.32) and (V.35), against time of fall T_f, with vertical speed V as parameter, emphasizes the practical utility of T_f as a major variable. The curves found within each family are very simple and very similar. In fact, if proper scales are chosen both complete families may be satisfactorily represented by a single master curve of the required correction against T_f. When this is done, the effect of V is merely to shift the origin with respect to which the master curve is used. The shift of origin along the T_f axis, that is, the climb-compensating increment to be added to T_f in using the master curve, varies linearly with V for both corrections, but at a different rate for each. The shift of origin along the $T'/(T_f+\tau_s)$ or S_0 axis respectively, that is, the climb-compensating increment to be added to $T'/(T_f+\tau_s)$ or to S_0 as determined from the master curve, varies in highly non-linear fashion with V.

Time of fall used must be that corresponding to the altitude A_r and vertical speed V_r at the instant of bomb release. Time of fall for the instant of release must already be computed and available for compensation at the instant that the Sniffer "sniffs", that is, accepts the slant-range and slant-speed data on which to base release. This instant precedes actual release by the time lag τ_s of the Sniffer and bomb-release mechanism. Altitude must actually be measured at an instant that precedes sniffing by the time lag τ_a of the altimeter and compensating servo,

at which time it has a value A_a. Vertical speed must be measured at an instant that precedes sniffing by the time lag τ_v of the speed-measuring device, at which time it has a value V_v. If V is determined from altitude data, the time lag of the speed-determining circuits and vertical-speed servo is $\tau_v - \tau_a$.

During each of the component time lags, vertical acceleration a alters both vertical speed and altitude, and vertical speed also alters altitude. Rate corrections of this sort must be brought to an end somewhere, and the assumption has been made throughout that the aircraft is flown with constant vertical acceleration during the entire short interval $\tau_v + \tau_s$ just before release. Then

$$V_r = V_v + a(\tau_v + \tau_s) \tag{VIII.12}$$

and

$$A_r = A_a + V_r(\tau_a + \tau_s) - \tfrac{1}{2}a(\tau_a + \tau_s)^2 \quad , \tag{VIII.13}$$

while in terms of T_f

$$A_r = \tfrac{1}{2}gT_f^2 - V_r T_f \quad . \tag{VIII.14}$$

From these relations

$$A_a = [\tfrac{1}{2}g(T_f - \tau_a - \tau_s) - V_r](T_f + \tau_a + \tau_s) + \tfrac{1}{2}(a+g)(\tau_a + \tau_s)^2. \tag{VIII.15}$$

Fig. VIII.-8 is the block schematic diagram of one alternative two-variable Sniffer compensator devised on these principles. The device is actuated by two servo motors, one setting a shaft to an angular position proportional to T_f and the other to V_v, with time-lag corrections determined by a vertical-axis accelerometer. The vertical-speed servo shown is of the simplest sort, producing by means of a linear potentiometer a follow-up voltage proportional to shaft position, for servo comparison with a data voltage proportional to vertical speed.

A more complicated arrangement is necessary for operation of the time-of-fall servo. Shaft rotation proportional to T_f is made, by a mechanism which builds up equation (VIII.15), to provide a follow-up voltage for servo comparison with a data voltage proportional in turn to A_a. Voltage from a regulated source is applied to a rheostat and potentiometer in series, and follow-up voltage for

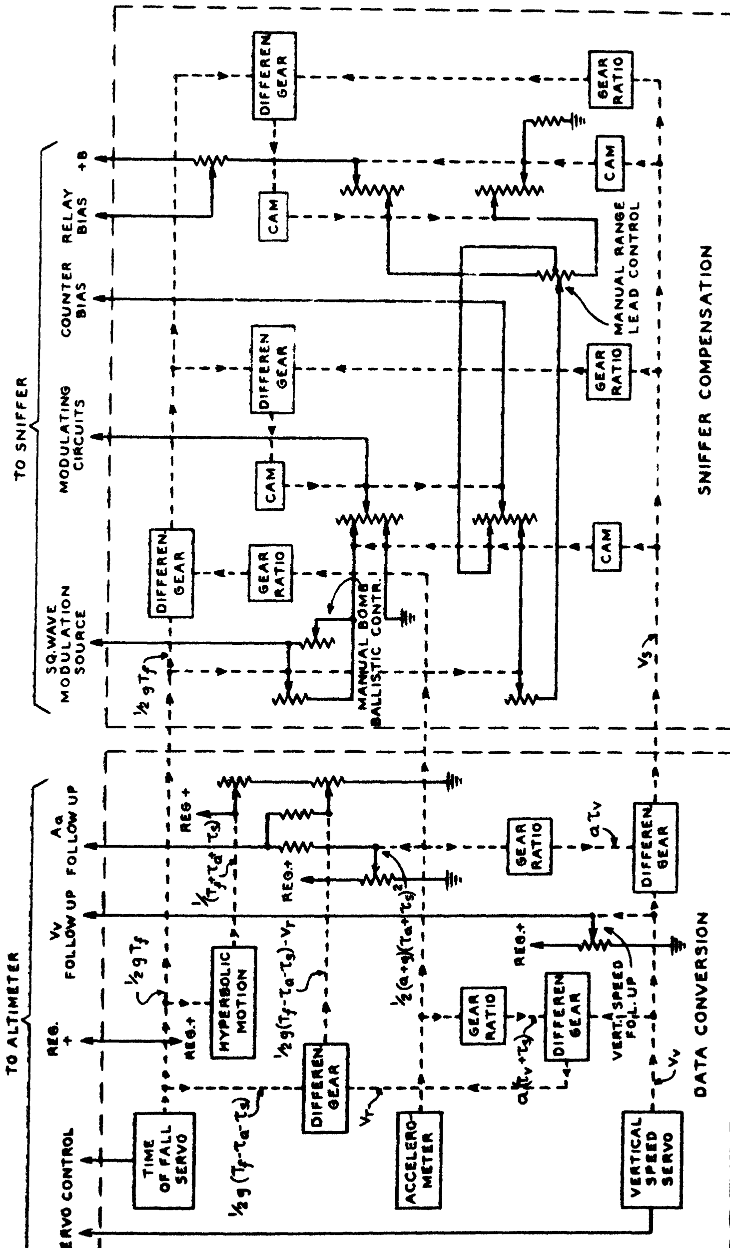

Fig. VIII.-8. Alternative two-variable compensator for bomb release from vertical maneuvers.

comparison with altitude-data voltage is taken from the arm of the potentiometer. The time of fall servo is connected to run in the direction to make this follow-up voltage equal the data voltage.

Assume that the servo shaft is at a position proportional to the time of fall T_f which is to be determined from altitude and vertical-speed data. By gearing out of the T_f shaft, either through a simple hyperbolic-motion mechanism to a linear series rheostat in the follow-up circuit, or directly to a series rheostat having a hyperbolic characteristic, the total resistance of that circuit can be made proportional to $1/(T_f+\tau_a+\tau_s)$. The voltage across a potentiometer in series with the rheostat is then proportional to $T_f+\tau_a+\tau_s$. The fixed time increment $\tau_a+\tau_s$ is merely a matter of proper choice of initial mechanical-position settings of geared shafts. By differential gearing of suitable ratios from speed-servo (V_v) and accelerometer (a) shafts, a shaft can be positioned in proportion to V_r [see (VIII.12)]. By a further differential from the time-servo shaft the position of the potentiometer shaft can be made proportional to $\frac{1}{2}g(T_f-\tau_a-\tau_s)-V_r$. The follow-up voltage at the potentiometer arm is therefore proportional to the product term of equation (VIII.15). The final term, $\frac{1}{2}(a+g)(\tau_a+\tau_s)^2$, is easily produced by a linear potentiometer controlled by the accelerometer and may be added electrically to produce an overall follow-up voltage proportional to A_a.

If the assumption of servo-shaft position proportional to T_f were not correct, the follow-up voltage produced as above would differ from the A_a data voltage; the servo would then run to nullify this difference and so set its shaft in proportion to T_f. A simple linear follow up from the time-of-fall shaft might be used instead if a time-of-fall data voltage were developed directly from the altitude and speed data, according to equation (V.37) as modified by acceleration. Since equation (VIII.15) is very much simpler in form, and T_f and V_v servos are necessarily to be available for use anyway, it is much easier to build up an A_a follow-up voltage as described. This inverse method of causing a servo to position a shaft in accordance with a prescribed function of observed data is of very

general utility, both in possible f-m radar applications and elsewhere. A hyperbolic rheostat in series with a linear potentiometer should be used to develop the required product, in preference to two linear potentiometers in cascade, because it avoids the distortion produced by loading a first potentiometer with a second one. In the arrangement shown, the adding circuit fed by the $\frac{1}{2}gT_f - V_r$ potentiometer must not load the latter appreciably.

After shaft positions representing time of fall and vertical speed are set up as described, they must be applied to compensate the f-m radar for bomb release. Time of fall may be used directly to set a linear rheostat in series with a potentiometer in the modulation-control circuit of the slant-range radar. There results a total circuit resistance proportional to $T_f + \tau_s$, hence a modulating-signal voltage across the potentiometer proportional to $1/(T_f + \tau_s)$. This is the primary range-scale control of the Sniffer being compensated. A similar control applied at high impedance to the output between arm and center of a relatively low-impedance manual range-lead potentiometer in the Sniffer bias-supply circuit provides corresponding compensation of range-lead scale. Range lead will be either positive (impact beyond target) or negative as the variable range-lead tap is respectively above or below the fixed center-tap.

Slope-correction factor $T'/(T_f + \tau_s)$ and speed intercept S_0 of the Sniffer range-speed approximation [see equations (V.32) and (V.35), using also (V.9)] are both functions only of the two variables $[A_s/(T_f + \tau_s)] + V_s$ and $A_s/[\sqrt{H_1 H_2}(T_f + \tau_s)]$, for any fixed horizontal speed-limit ratio H_2/H_1. In view of the effects of time lags in equipment, examination of these variables as expressed in terms of T_f, V_s, and a indicates that time of fall used in computing corrections should be modified slightly from the value at release. For time-of-fall limits of 2.8 to 12 seconds, with τ_s at 0.4 second, the modified time required is always very close to $T_f - 0.35 - 0.80\, a/g$ seconds. For vertical speed, only the value V_s at the actual moment of sniffing is required. Time and speed servo outputs T_f and V_v are easily modified to these values by data from the accelerometer, with the aid of differential gearing.

Increments proportional to V_s by the proper two factors are added by further differential gearing to time of fall as corrected for acceleration. The resulting shaft motions act through two cams shaped in accordance with the master correction curve, one to move the stator windings of linear correction-factor potentiometers in the modulation circuit and in the range-lead circuit, the other to move the windings of twin linear rheostats in the bias circuit. The V_s shaft also acts alone, through two cams shaped in accordance with required correction increments, one moving the arms of the modulation and range-lead-scale correcting potentiometers and the other the arms of the bias-circuit rheostats. Connected to maintain constant total resistance, the twin bias-correcting rheostats have the manual range-lead potentiometer between them and together act as a split potentiometer to set the Sniffer bias with zero range lead.

By means of these linear electrical elements with double non-linear mechanical inputs, the correction factor $T'/(T_f+\tau_s)$ is applied to the Sniffer modulation sweep, and bias corresponding to speed intercept S_0 is applied to the Sniffer-counter load. Alternatively, double linear mechanical drives may be applied to chains of non-linear rheostats and potentiometers to achieve the same result without the use of cams.

As in the compensator using a three-dimensional cam, rough correction for bomb trail can be applied by shunting a manual rheostat across the T_f rheostat of the sweep-control circuit. Similar correction of range-lead scale would give proper tracking of controls and could easily be provided. Means for making still more accurate allowance for trail, including its slight variation with V, were not worked out.

Of the two compensators described and quantitatively worked out, only that using the solid cam was built. Either would have worked with quite normal Sniffer radar and counter equipment. The one constructed was markedly the better in mechanical simplicity, and to some extent in electrical simplicity also, though there was no major difference in the latter respect. The composite compensator using many simple elements would be much more easily altered to meet varying requirements, and because of its

flexible design permits the more accurate compensation for effects of vertical acceleration. Studies of trail correction were not carried far enough to give a final comparison, but as far as development was carried for the two methods, an advantage for the solid cam was indicated. Both have been described as representative of possible ways to apply the type of data provided by f-m radar to somewhat more complex fire-control problems than that of level-flight bombing.

e. ***Exact Solution.*** All Sniffer bombing computers so far described have given solutions based on linear approximation of the quadratic relation among slant range R, slant speed S, altitude A, and vertical speed V at any given time T prior to passage over the target. The exact relation [see equation (V.30)] is

$$^{R}/_{T}(^{R}/_{T}-S)-{^{A}/_{T}}(^{A}/_{T}+V)=0\ . \tag{VIII.16}$$

Overall range sensitivities of altimeter and Sniffer may be kept equal, while both are varied in inverse proportion to a shaft position which may be assumed to represent time T to passage over target. Neglecting or compensating cathode-follower imperfections, a Sniffer using both simple and switched counters may then be made to produce separate output currents representing R/T and $^{R}/_{T}-S$. An altimeter with similar counters may give output currents A/T and $^{A}/_{T}+V$. With such data, an automatic solution exact for any values of the variables may be obtained by very simple means.

A relay of the type indicated in Fig. VIII.-9 is the only special element required. Two electro-dynamometer instrument movements mounted on a single shaft, as in a two-phase wattmeter, are required. The shaft must also mount as shown a contact finger or other means of effecting relay action. Currents representing respectively R/T and $^{R}/_{T}-S$ in the two separate coil circuits of one dynamometer movement will produce a torque on the common shaft proportional to the product $^{R}/_{T}(^{R}/_{T}-S)$. Similarly, currents of suitable polarity in the other movement, representing A/T and $^{A}/_{T}+V$, will produce an oppositely directed shaft torque proportional to the product $^{A}/_{T}(^{A}/_{T}+V)$.

If T is true time to target as assumed above, these torques will just cancel as in equation (VIII.16), the shaft will not move, and the relay will not be made to close on either contact. If the range-sensitivity setting existing in altimeter and Sniffer does not exactly correspond to time until target crossing, one torque will predominate, the common instrument shaft will turn, and the relay will close on one contact. Closure of either contact may actuate a reversible servo motor in the proper direction to drive the range-sensitivity control to a setting truly representing time to target. Automatic solution for the exact value of T is thus accomplished.

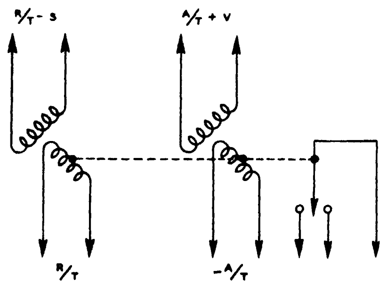

Fig. VIII.-9. Product-comparing relay.

The product-comparing relay of Fig. VIII.-9 operates by cancellation of two electrodynamic torques, so requires no spring torque on the common shaft. The shaft never rotates by any significant amount, so variation with angular shaft position of the torque characteristic of the dynamometer movements is negligible. This special relay would therefore be inherently a stable and reliably accurate instrument.

For aircraft motion in a straight line in the vertical plane through the target, a tracking clock used as described in section 5b of this chapter is a useful addition to the exact-solution Sniffer system. Even when moderate vertical maneuvering is done, the time to target is not greatly altered except in the normal way by forward motion of the aircraft relative to the target, so the tracker retains considerable usefulness. If time lags in establishing the four outputs R/T, $R/T - S$, A/T, and $A/T + V$ can be equalized, no corrections for vertical acceleration are

needed. Otherwise, such corrections need only set up the four data elements for some convenient common instant.

No bias corrections to the Sniffer are required for the simple exact computation. If, however, residual range R_r is present, or range lead R_l is required, corresponding scale-corrected bias increments are necessary to make the Sniffer output currents become respectively $R/T + (R_l - R_r)/T$ and $R/T + (R_l - R_r)/T - S$. Similar bias correction for residual altitude of the altimeter installation is necessary also.

Bomb release may be accomplished quite conveniently by use of the time-to-target shaft position, though allowance for time lag then becomes necessary. Neglecting time lags, equation (VIII.14) indicates that

$$A/T + V - \tfrac{1}{2} g T = [\tfrac{1}{2} g(T_f + T) - V](T_f - T)/T \quad . \qquad \text{(VIII.17)}$$

Release should take place just when time to target becomes equal to time of fall, making the right-hand side of (VIII.17) zero. That is, release should be made to occur when data-output voltage $A/T + V$ just equals a voltage $\tfrac{1}{2} g T$ derived by a linear potentiometer from T-shaft position.

Actually, the T of equation (VIII.17) must be time from target at release, T_r. Data voltage $A/T + V$ is, however, that for an instant of altitude determination earlier than T_r by a total time lag $\tau_a + \tau_b$, in altimeter and bomb-release mechanism. Data voltage available is therefore $A_a/T_a + V_a$, and becomes available to initiate release at a time to target T which is $T_a - \tau_a$. Release will occur at a time to target T_r which is $T - \tau_b$ if release-mechanism action is iniated by voltage equality at time T. Allowing exactly for these time lags in the case of constant vertical acceleration a, correct release will occur if initiated at actual time to target T by the voltage relation

$$A_a/T_a + V_a - \tfrac{1}{2} g (T - \tau_b)^2/T + \tau_a$$
$$+ a(\tau_a + \tau_b) - \tfrac{1}{2} a \, \tau_a^2 - \tau_b^2/T + \tau_a = 0 \quad . \qquad \text{(VIII.18)}$$

There is, of course, no difficulty in determining $T - \tau_b$ from a clock shaft which indicates T.

Aside from the time-lag corrections, then, automatic

bomb release by this computing system merely requires comparison of time-shaft position with one data output of the altimeter. Release is computed exactly (for vacuum fall) if good radar data is setting the time shaft at and just prior to release. Lag corrections require merely the use of a slightly non-linear time-shaft output, as well as the addition of a small time-weighted acceleration term to the condition for release.

Corrections have been discussed at some length in the case of all the vertical-maneuvering computers because they are responsible for much complication of basically simple fire-control systems using f-m radar. A large part of system-development planning must therefore be devoted to finding methods of correction which cause a minimum of complication. Fig. VIII.-10 is a block diagram of a complete bombing system of the type described. Means for applying bomb-trail correction have not been worked out for the exact computer, but should not prove too difficult if required.

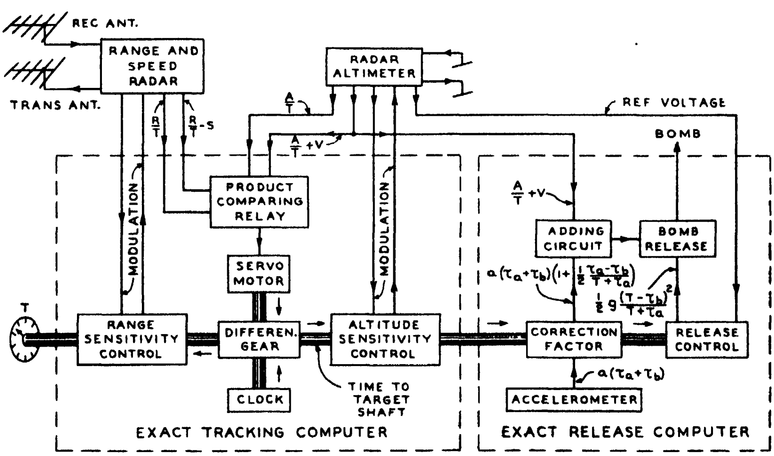

Fig. VIII.-10. Exact computer for bombing from vertical maneuvers.

Errors caused by approximations used in the other computers described are not serious in comparison to errors usually characteristic of f-m radar and counter operation. Samples of product-comparing relays which were tested proved too insensitive for use with tubes drawing acceptably low plate current, and too susceptible to vibration

for satisfactory operation in aircraft. These difficulties might have been overcome, but actual needs did not justify the necessary effort. The ease with which the single-servo exact computer lends itself to clock tracking, as well as its general simplicity, does make the system attractive. Clock tracking might do much to render usable the impaired radar signals expected in maneuvering flight.

8. OPERATION AT 4000 MEGACYCLES

a. *Purpose and Results.* It was desired that f-m radar equipment be available for operation on still a third frequency, in addition to the equipment developed for 410 and 1500 megacycles per second. From available 4000- and 6000-megacycle frequency bands, 4000 megacycles per second was chosen as center frequency for further work. The usefulness of higher antenna directivity and higher power than had been available previously was to be tested at the new frequency. A transmitter power of 20 watts was chosen.

Experimental work at 4000 megacycles[2] led to successful development of f-m radar techniques for that frequency and to understanding of problems involved, but did not reach the stage of construction of service-prototype equipment. Radio-frequency components had to be specially built for the uncommon frequency used, and transmitter tubes and methods of frequency modulation had to be developed from the very beginning. An urgent requirement that some equipment be in airborne operation as early as possible prevented extensive revision of the original experimental apparatus, when experience accumulated with it had indicated the need of such measures.

When stopped after the close of the war, the 4000-megacycle system work had reached a point at which design of prototype equipment could have been undertaken. Signals had been observed in a monitoring oscilloscope on a number of test flights and range-counter operation had been obtained, though without accurate calibration. Maximum ranges of three to four miles attained on targets of moderate size were comparable to results with the *AN/APG-6(XN)* equipment and, though no direct comparisons were made, were probably well in excess of ranges attainable with the *AN/APG-4* or *AN/APG-17* equipments. This good result indicates that the

high antenna directivity and high transmitter power were being used to advantage. Preliminary flight tests without modulation gave indications that ground speed of aircraft can be determined by use of the Doppler frequency shift of the reflected signal. The flight tests also showed that extreme care in reducing feed-through noise would be necessary to permit effective use of substantially higher powers at such frequencies.

The other systems described in this chapter show how f-m radar data can be applied to solution of special problems, and call for only minor modifications of the radar equipment itself. The 4000-megacycle work, on the other hand, developed a new f-m radar data source without investigating special applications of the data.

b. *General Description of Apparatus.* Preliminary work at 3000 megacycles indicated that balanced radio-frequency detectors can be made at such frequencies, by using crystals in carefully built and adjusted transmission-line circuits. However, balanced detectors have been found to require a multiplicity of critical adjustments and to exhibit inadequate stability. Only superheterodyne methods, not requiring balanced detection at radio frequency, were considered for use in the 20-watt, 4000-megacycle system. Because of the extreme sharpness of filter characteristic required for its operation at intermediate frequencies then efficiently usable, the side-band superheterodyne described in section 5b of Chapter III. and used in *AN/APG-6* and *AN/APG-17* equipments was not considered applicable.

Effort was concentrated on development of a signal-following superheterodyne, using the principles described in section 5c of Chapter III. and laid out in accordance with the block diagram of Fig. III.-26. Development of this alternative method of operation and study of its peculiarities was considered desirable in itself. This system requires automatic frequency control of a local heterodyne oscillator, so as to follow the frequency modulation of the radar transmitter. Control is obtained by use of error signal from an intermediate-frequency discriminator to maintain the beat between transmitter and local oscillator at a substantially constant frequency. Radar information is taken from the received signal, at

intermediate frequency, by beating it with the control-channel i-f signal in a balanced second detector. This process cancels the residual frequency modulation present on both channels because of imperfect frequency control, leaving only the desired range and speed beat frequencies.

A continuous-wave vane-type multicavity magnetron with a permanent-magnet field is used in the experimental equipment as transmitting oscillator, and a reflex klystron as signal-following local oscillator. Electronic regulation of the magnetron current is used to ensure power-input stability with changing load on the magnetron. Because of the need to operate oscillator and regulator cathodes far from ground potential and to supply several separate high voltages, the system is powered from a source of 800-cycle alternating current rather than from storage batteries.

Several ways of frequency modulating the transmitting oscillator were investigated, and none tried was found fully satisfactory. Variable reactances coupled to the magnetron separately from the output load,[3] and operating as described in section 4d of Chapter III., were tried. Study of circuit conditions indicated that minimum amplitude modulation and maximum frequency modulation are obtained by connecting the variable reactance either at a maximum-impedance (anti-resonant) point or at a minimum-impedance (resonant) point of the coupling circuit to the magnetron. An impedance transformation between the variable reactor and the coupling circuit is often advantageous. Diode-loaded resonant lines operated well as variable reactors, but diode life was excessively short because of heavy cathode bombardment, modulation characteristics depended upon magnetron loading, and adjustment for linear modulation was critical and not fully stable. Rotary capacitors were considered but not used, because they are not suitable for control by electrical means of the modulation sweep and thereby of range sensitivity. Vibrating capacitors were found to spark over or to change calibration, because of radio-frequency heating of the diaphragm at the level of transmitter power in use.

It is probable that a satisfactory modulator using an external variable reactance could have been developed.

However, magnetrons then available with single output-coupling loop were not suitable for such service and the supply of tubes with two separate coupling loops was not adequate for system development. No additional dual-loop tubes could be obtained because effort in the tube-development phase of the f-m radar program was concentrated on internal modulation by helical electron beams. The internally modulated tubes, discussed in section 4e of Chapter III., would no doubt have been very useful, but were not available in time for the 4000-megacycle system tests.

All system tests were made with magnetrons modulated in frequency by variation of anode current. Use of an electronic current regulator in the magnetron power supply facilitates such modulation by current variation, since the operating point of the regulator can be controlled by the modulating signal. Sweep widths up to 5 megacycles per second are obtained, at a modulation frequency of 400 cycles per second. This method of modulation proved very simple but not fully satisfactory.

Fig. VIII.-11 is a functional circuit diagram of the experimental equipment. Values of circuit elements, given as actually used, are illustrative and do not necessarily represent recommended design. The radar-beat amplifier most used is of the typical sloping-characteristic type, with selective feed back applied to the first stage and peak response at 25 kilocycles per second. It is not shown in the circuit diagram, nor is the conventional single limiter and meter-output range counter of the type shown in Fig. IV.-2. Power supplies are also conventional and are not shown.

The reflex-klystron local oscillator, similar to the *723A/B* but adapted for 4000-megacycle operation, is of course pretuned by mechanical deformation of its cavity and frequency modulated by control of its reflector voltage. A separate electronically regulated power supply with positive terminal grounded is provided for the klystron oscillator. Regulated plate voltage is also used for the d-c amplifier applying automatic frequency control to the reflector electrode of the klystron, and is obtained from the receiver power supply. A separate control-channel intermediate-frequency output stage supplying the a-f-c

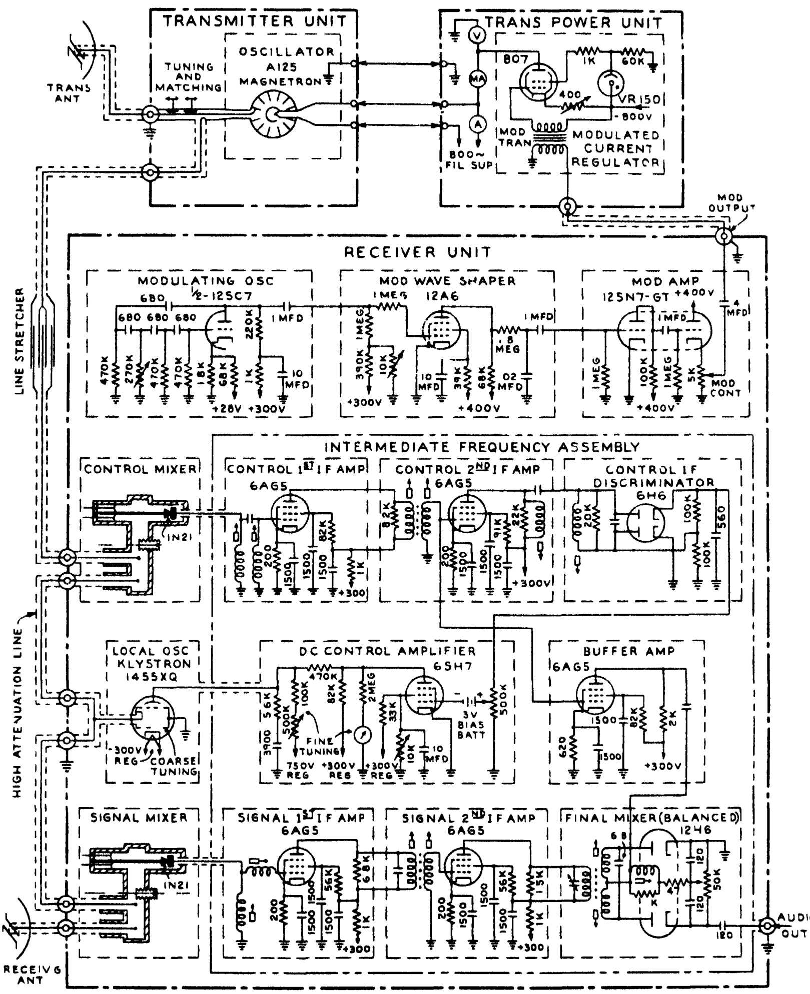

Fig. VIII.-11. **Functional circuit diagram of experimental 4000-megacycle f-m radar.**

discriminator will be noted in the circuit diagram; this is required to overcome tuning interaction which was observed between discriminator and balanced second detector when both were driven by a common source.

c. *Difficulties and Solutions.* Most of many difficulties encountered in getting the new system into operation fell in one broad category: noise. Noise, mostly not

of random nature, arose primarily in two ways: by imperfections of power sources and tubes, and by spurious radar signals resulting from feed through or cross coupling between transmitter and receiver. Feed-through effects could be found at both the first and second detectors of the signal channel. Similar difficulties of course had to be overcome in carrying out the development of the earlier lower-frequency systems also. Microphonics provided ever-present difficulties in this as in all other airborne equipment. Use at 4000 megacycles of much higher transmitter power than had been used at lower frequencies greatly accentuated the noise problems.

One source of noise lies in the fields of the a-c operated heaters of the magnetron-current regulator tube, as well as those of the magnetron and klystron tubes; none of these heaters can be grounded. This noise is aggravated by rectifier-starting pulses fed back to the power circuits from the magnetron power supply. A number of ordinary measures together proved able to reduce these difficulties to an acceptable level, though they were never fully eliminated. Other pulse noise seems related to possible sharp imperfections in the frequency characteristic of the a-f-c discriminator. Still other pulse noise results from small, sharp oscillation-amplitude discontinuities found to occur when frequency modulating magnetrons. This is a most serious source of noise. A final source of excessive noise from circuits and tubes is the magnetron noise described in section 3b of Chapter III. This has been found in a special study[4] to be caused by condensible vapors in the magnetron cavity, and is kept at an acceptable level by operating the magnetron at somewhat reduced voltage.

The balanced second detector or final mixer shown in the circuit of Fig. VIII.-11 is used to prevent noise from appearing as a result of stray amplitude modulation of magnetron or local oscillator. Amplitude modulation acting on either the signal or control channel alone should be balanced out of the second-detector output. Amplitude modulation on a strong control-channel signal in the presence of a weak unmodulated signal-channel signal should also fail to produce unwanted output, using diodes as

linear detectors. Suppression of control-channel modulation was never found to be complete, however, because of cross coupling from the control-channel input of the parallelled diodes to the signal-channel input of the diodes in push pull. Great care in the design and construction of the balanced detector-input transformer reduces this cross coupling to a tolerable level. Experiments with pentagrid mixers rather than diodes in the balanced circuit indicate that still greater freedom from cross coupling could be achieved by using such tubes.

Size and shape of the two relatively highly directive antennas, each using paired dipole radiators in a paraboloidal reflector of 16-inch mouth diameter, necessitates their installation side by side in the nose of the aircraft. This close proximity makes reduction of feed through, or cross coupling, directly between antennas a problem of major difficulty. A number of measures to reduce such coupling and its effects must be used together. Quarter-wave choke grooves around the reflector circumferences reduce coupling appreciably with the antenna systems mounted in the open, as does the use of spaced-dipole antennas each producing a direct-radiation pattern null in the direction of the other antenna.

Mounted side by side in the clear plastic nose of an aircraft, which serves as a radome, the antennas are separated by a vertical plane metal sheet or septum. This septum is cut to fit the contour of the radome so as to form as complete a partition as possible between the antennas. Energy reflected by the radome in front of the transmitting antenna can still reach the region behind the antennas, however, and if there again reflected will reach the radome in front of the receiving antenna, where it will be reflected still a third time and so reach the receiving antenna. To minimize such very troublesome triple-reflection feed through, a surface of special nonreflecting material (supplied by the Radiation Laboratory of the M.I.T.) is set up behind the antennas.

Still another source of head-end feed through is evident from the circuit diagram, Fig. VIII.-11. Signal from the transmitter, coupled to the control-channel mixer or first detector, will enter the line connecting this mixer to the

local oscillator. Through the common local-oscillator coupling, it will then reach the connecting line to the signal-channel mixer or first detector, appearing on that mixer as a feed-through signal. Feed through of this sort is reduced by running the local oscillator at a high output level and using strongly attenuating connections (100 db. loss) from it to the two first detectors

Feed through is troublesome because it results in the appearance of a frequency-modulated signal, just like the desired target-reflected radar signals, at just the places in the circuit where the desired signals appear. The spurious signal is likely to be much stronger than the desired signal and can therefore cause a lot of trouble. The measures already described reduce actual unwanted-signal cross couplings. They must be used in conjunction with other measures serving to reduce the damage done by such unwanted signals as remain.

Local mixing signal arrives at the balanced second detector or final mixer, at intermediate frequency, over a path of total length determined by the electrical length L_1 of the connection from transmitter to control-channel first mixer and length L_2 from local oscillator to that mixer, as well as the equivalent electrical length L_3 of the control-channel intermediate-frequency amplifier. These path components are shown in Fig. VIII.-12. Similarly,

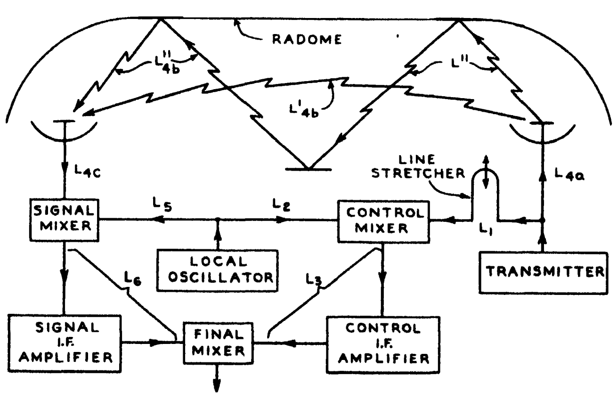

Fig. VIII.-12. Cross-coupling paths.

feed-through signal reaches the second detector over a path determined by electrical length L_4 of the undesired

cross-coupling path from transmitter to signal-channel first mixer and length L_5 of the connection from local oscillator to that mixer, as well as the equivalent length L_6 of the signal-channel intermediate-frequency amplifier.

Equivalent electrical length of a selective intermediate-frequency amplifier is proportional within the pass band to the slope $d\psi/d\omega$ of the phase-frequency characteristic of the amplifier. This is the same property used in the artificial calibrating delay lines described in section 3b of Chapter VII., the phase slope measuring directly the time delay of a signal front in passing through the selective circuit. Frequency-modulation sweep of the intermediate-frequency signals is reduced to a small fraction k of that of the transmitter by automatic frequency control of the local oscillator. Path length in the i-f channels is therefore correspondingly reduced in its effects as compared with equal circuit length in r-f portions of the system. Large equivalent path length can easily occur in a poorly aligned i-f amplifier, however. Reduction by a factor k as low as 0.02 is found possible in the experimental system.

The effect of mixing a local reference signal with an incoming signal, whether from a distant target or through a local cross-coupling path, of course depends both on the total difference in length of the paths by which the two signals reach the final mixer and on the time rate of change of transmitted frequency. If the path difference is great, the two frequency-modulated signals will have distinctly different frequencies when they reach the mixing point and a corresponding range-beat frequency will appear in the second-detector output. If the path lengths differ only slightly, the frequency difference will be so slight that only a portion of a beat cycle occurs during a single modulation sweep of transmitter frequency. In this case the fractional beats at the second-detector output can only recur at the modulation frequency (or at twice that frequency for certain r-f phase conditions), as described in section 2g of Chapter IV. and shown in Fig. IV.-11. This is a common condition of operation, and is likely to result in overloading of the beat-frequency amplifier if large cross coupling is present.

If, finally, the two paths have exactly the same length, the two signals being mixed will maintain exactly equal frequencies, and will behave exactly alike as their common frequency is swept in modulation. They will therefore produce no beat-frequency or modulation-frequency output at all from the second detector under this condition, and cross coupling which meets the condition will do no harm if it does not overload the mixers. The condition of equal total paths requires that the overall path difference

$$d = (L_4 - L_1) - (L_5 - L_2) + k(L_6 - L_3) \qquad \text{(VIII.19)}$$

shall be zero, where k is the factor W_{i-f}/W by which a-f-c reduces the frequency swing of the i-f signals.

It is impracticable to use delay differences between i-f amplifiers to cancel r-f delay differences, since the two delays are not likely to behave similarly under various disturbing influences. L_3 and L_6 must therefore be equalized by matching phase-characteristic slopes over the common pass band of the two amplifiers. Cross coupling by leakage from transmitter to signal detector through the control detector and the local-oscillator connections provides a path L_4 of length $L_1 + L_2 + L_5$. Total path difference d for this case with i-f amplifiers equalized is therefore given by (VIII.19) as $2L_2$. The connection from local oscillator to control-channel first detector must therefore be made of negligible length, while both local-oscillator connections retain high attenuation, in order that cross coupling through these connections will both be minimized and rendered harmless.

Cross coupling may also occur over a path external to the antennas, for example L'_{4b} directly between antennas, or L''_{4b} by way of triple reflection from a radome before the antennas and some other object behind them. This case involves a path L_4 including also the electrical lengths L_{4a} of the connection from transmitter to transmitting antenna and L_{4e} of the connection from receiving antenna to signal-channel first detector. If the two i-f amplifiers are equalized and local-oscillator connectors L_2 and L_5 are both reduced to negligible length or otherwise made equal, total path difference d can be reduced to zero by adjusting the length L_1 of the connection from transmitter

to control-channel first detector so as to equal the total length L_4 from transmitter to signal-channel detector. This adjustment, referred to as a "line stretcher" in Figs. VIII.-11 and VIII.-12, can of course only be made exactly for a single cross-coupling path at a time. If several paths of distinct lengths are present at once, only a compromise adjustment can be made.

Spurious modulation of the transmitted signal by microphonics, power-supply hum or other stray disturbances usually includes frequency modulation as well as amplitude modulation. Repetition frequencies of such frequency modulation usually lie in a useful range of radar-beat frequencies. Cross coupling with a small path difference will then produce second-detector output at the frequencies of the spurious frequency modulation. Because cross coupling is likely to be strong at best, and the spurious frequencies to have troublesome values, marked disturbance of desired operation results if path lengths are not accurately matched. This is especially true when operation at very short ranges is required as in the case of an altimeter, the cross-coupled signals then becoming practically indistinguishable from the desired reflection unless the total cross-coupling path is accurately equal to the total local mixing-signal path.

To summarize, the following measures have been found useful to ensure good signal/noise ratio in frequency-modulated radar equipment using a signal-following superheterodyne receiver:

(1) Minimize electrical and mechanical disturbances, especially to transmitter and local oscillator and to objects in the field of the antennas.

(2) Minimize modulation of the oscillators by residual disturbances.

(3) Minimize effect of amplitude modulation of either oscillator by good balance of final mixer, by avoiding cross coupling of the two input channels to this mixer, and by using relatively strong control-channel input to the mixer.

(4) Minimize strength of cross couplings in r-f portions of system by attenuation in local-

oscillator connections and by care in location and treatment of the antennas and their surroundings.

(5) Minimize harmful effect of frequency modulation of cross-coupled signal by equalizing its path with that of the local mixing signal, thus:

(a) Equalize slope of phase-frequency characteristics of the two i-f amplifiers.

(b) Minimize and equalize lengths of local-oscillator connections.

(c) Minimize length of cross-coupling path, and adjust connection from transmitter to control-channel detector to have equal length.

Much of the preceding discussion applies to all f-m radar systems. It is emphasized here because cross coupling was particularly serious with the adjacently mounted antennas of the 4000-megacycle equipment and because a particularly thorough investigation of the sources of disturbances had to be made in the case of that equipment, in order to realize the good performance made possible by the relatively high power used. Careful use of the expedients listed above resulted in successful operation of the 4000-megacycle system with signal-following superheterodyne receiver. The expedient of equalizing coupling paths should also aid in operation of altimeters at very short ranges and with antennas mounted close together.

9. SELF CALIBRATION IN RANGE

Ranging accuracy of frequency-modulated radar is primarily controlled by accuracy of modulation. The simple averaging range counter is itself capable of great accuracy, particularly when used as a null device (see sections 2d and 5c of Chapter IV.). Accurate, stable control of frequency-modulation sweep width is on the other hand quite difficult, especially with a vibrating modulator. It is likely that the situation could be improved by applying inverse feed back to the modulator-driving circuits, with the modulator-diaphragm motion included in the feed-back loop; this has not been tried.

Another method often suggested but probably never tried for the improvement of ranging accuracy is inclusion as a

permanent part of the system of a known propagation-delay element to make the equipment self calibrating. Very good and stable accuracy should be attainable in this way, since the f-m radar system would act merely as a neutral device comparing the unknown target range to the known calibrating range. Two range-beat frequencies would be produced by the radar, with the frequencies remaining in the same ratio as the ranges to be compared, independently of radar range sensitivity. The counters or other data-comparing system would merely have to determine the frequency ratio, independently of the absolute values of the frequencies or the counter sensitivity.

Self-calibrating systems using actual transmission lines or artificial filter-type lines are unattractive on the score of weight, bulk and attainable calibrating range. These objections do not apply, however, to the multiple-beat types of artificial calibrator described in section 3c of Chapter VII., and wherever the need for high accuracy justifies the extra complication an entirely practical self-calibrating system should be possible.

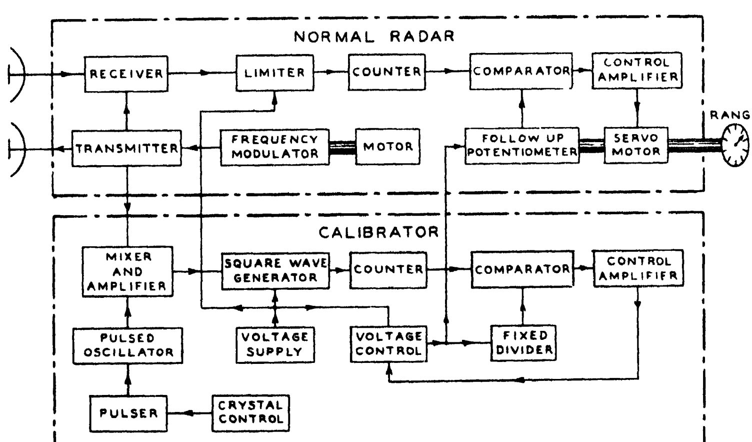

Fig. VIII.-13. Self-calibrating ranging radar.

Fig. VIII.-13 indicates one way in which such a system might be organized. A normal f-m ranging radar with a motor-driven modulator and a servo-balanced null counter is indicated. The additional calibrating channel is based on a source of many equally spaced fixed frequencies in the

radar operating band; this source may be an oscillator sharply pulsed at a crystal-controlled pulse frequency. From the beats between the sweeping radar signal and the "picket fence" of calibrating frequencies, a calibrating square wave of typical range-beat frequency is developed. This square wave in the calibrating channel is applied to a second null-type counter, working against a follow-up voltage derived by a fixed divider from the same source that supplies the servo-driven follow-up potentiometer of the normal radar channel. The calibrating-channel counter is balanced by automatic control of the total voltage applied to the two follow-up dividers.

Calibrating range is determined solely by the crystal-controlled pulse frequency, so may be very accurately maintained. The limiters supplying square waves to the counters, the counters themselves, and the counter-null comparator circuits must accurately maintain a known ratio of counter sensitivities in the two channels. The voltage-division ratio of the servo follow-up potentiometer in the range channel will then be to the fixed division ratio in the calibrating channel as the actual target range is to the fixed calibrating range, so long as the automatic null balance is maintained in both channels. Both channels are fully compensated against supply-voltage variations.

Alternatively, automatic control of the follow-up supply voltage may be omitted if the radar is modulated by the usual vibrating capacitor. The calibrating channel may then be balanced by applying automatic control to a simple sweep-width compressor such as that used in the *AN/APG-6* and described in section 4b of Chapter VI. Fixed-error effects in the radar channel will normally be averaged out by small random variations of effective target range. This will not occur in the calibrating channel, however, and some artifice to insure averaging will be required. One way to meet such a requirement is to modulate slowly the frequency of the pulsed oscillator, so that the part of its picket-fence spectrum of equal-strength side frequencies lying in the modulation band of the radar will move back and forth rather than remaining fixed.

Many other self-calibrating arrangements are of course possible, but those suggested should suffice to indicate

what can be done. Radar speed sensitivity depends only on radio carrier frequency and that frequency can be very accurately maintained. Self calibration for speed is therefore inherent in the radar and requires no special facilities. Maintenance of calibration of switched speed counters is, however, quite another matter.

10. General Purpose System

There is sometimes need for direct indication of slant range to and slant rate of closing on an isolated target. The various special-purpose systems discussed have provided just this data but have utilized it in an internal computer rather than simply indicating it. Requirements of particular computers have in some cases conflicted with optimum radar design for simple range measurement. A radar system providing both range and speed data directly in the form of servo-shaft positions would serve generally for a wide variety of special purposes. It could be designed specifically for optimum data determination over the widest possible variations of range and speed. Its outputs could be indicated directly or could actuate a computer for any special purpose.

A general-purpose f-m radar data-gathering system with servo outputs was included in the development program, but because of urgent requirements for completion of specialized systems only limited preliminary study could be given the more general problem before the close of the war terminated the program. A general-purpose system, because of its wide usefulness, would justify much more effort toward refinement of design, both as to principles and as to practical details of manufacture and maintenance, than did the various special systems.

A fully integrated system having shaft outputs proportional to slant range, slant speed, target azimuth, aircraft altitude, and vertical speed, preferably with acceleration components measured to permit limited data prediction, would have very wide applicability indeed to problems of fire control against isolated targets. Results obtained with the more limited actual equipment give basis for belief that such an integrated system is attainable within size and weight limits tolerable for use in small

military aircraft.

Since no final planning was done on general-purpose slant-range and slant-speed radar, no more than a few suggestions resulting from preliminary study can be given here. The problems are to provide for operation from minimum ranges of a few hundred feet or even less to maximum ranges of several miles, with good and easily maintained accuracy, and to cover a closing-speed range from about 150 to perhaps 2000 feet per second with good accuracy and negligible disturbance by the range signal. Because of operating limitations of individual aircraft, it should seldom be necessary to cover even a large fraction of the entire range under any one condition of adjustment.

For reliable ranges of several miles on targets of moderate size, experience has indicated that a number of watts of transmitter power and a receiving antenna of appreciable effective area are required. R-f transmission-line losses should be kept low, for example by placing all r-f circuits directly at the antennas. Experience has also indicated that the complication of superheterodyne systems is justified by their reliability of operation.

Variation of modulation frequency or sweep width is necessary to provide best accuracy over the widest limits of range. There is much room for ingenuity in putting these variations to best use. Satisfactory sweep control and radar operation are obtainable over at least a 5:1 and perhaps a 10:1 range of sweep width; modulation-frequency control has been less convenient and therefore less investigated. Where small ratios of maximum to minimum range are acceptable, there is much to be gained in discrimination against noise by using a narrow beat-frequency band. In that case the range servo may be made to adjust sweep width, and perhaps also modulation frequency, so as to maintain a constant range-beat frequency with varying range and thus permit an amplifier band only wide enough to pass all speed frequencies.

Accurate range measurement is possible with range-beat frequencies covering a band as much as eight to ten octaves wide. The full band may be required in altimeters, where extremely low minimum ranges must be measured. By combining moderate variation of radar range sensitivity with a

counter operating over a moderate frequency band, widely varying range data should be determinable to good accuracy and a reasonably good signal/noise ratio should be obtained as well. For example, the range-servo controlled unbalance of a null-type counter might be arranged to adjust sweep width in accordance with reciprocal cube root of indicated range, and to adjust balancing follow-up voltage to the counter in accordance with the square of the cube root of range. Operation from 200 feet to 5 miles would then be obtained with a 5:1 variation of sweep and a 4½-octave beat-frequency band; the gain-frequency characteristic of the beat-frequency amplifier would be arranged to rise at about 8 decibels per octave.

The problem in speed measurement is to prevent large range frequencies from disturbing the measurement of small speed frequencies, as a result of inadequate switched-counter balance or some equivalent imperfection. It is very difficult to avoid having a large range frequency together with a small speed frequency under some condition of operation, if wide variations in both range and speed are to be indicated. Alternate operation of the radar with and without frequency modulation, to permit alternate measurement of range only and of speed only, has been suggested. Experience with residual stray modulation, particularly in the development of the *AN/SPN-2*, has indicated that this might be difficult, but it is certainly a possible solution.

Another type of modulation control that has been suggested and might repay development requires that the product of modulation frequency and sweep width be made to vary linearly with indicated speed and reciprocally with indicated range. This system has the interesting property that range frequency is maintained in constant ratio to speed frequency, and radar beat frequency during the frequency-modulation downsweep in fixed ratio to that during the upsweep. The possibility of limiting fractional speed error due to counter unbalance therefore exists, and requires no particular complication, but appears also to limit unduly the variations in range and speed that can be handled.

Had it been developed, the general-purpose servo-

output range and speed radar would have been the logical culmination of the single-target portion of the f-m radar development program. These suggestions as to paths that the development might have taken seem a fitting ending to the present discussion of special developments proposed or accomplished.

11. SYSTEMS RELATED TO F-M RADAR

Principles and techniques typical of frequency-modulated radar are also applicable to other systems which are related or similar to but are not truly f-m radar. A few of the related systems so far proposed or developed may be mentioned here.

Continuous-wave measurement of speed only by Doppler effect is typified by the *AN/SPN-2* system described in section 7 of Chapter VI. Other c-w systems, in which the Doppler beat is presented aurally and speed is not actually measured, have also been quite extensively developed and used.[5] These systems make use of the remarkable capabilities of the human ear for detecting and interpreting weak signals in the presence of noise. Such systems are particularly valuable where a small moving target must be detected in the presence of many large fixed reflectors, for example in the detection of an airplane flying among mountains or of a soldier moving in a forest.

In difficult detection problems such as those just mentioned, determination of range is desirable but of secondary importance. Some range information is obtainable by multiple-frequency continuous-wave operation,[5] target ranges being determined by relative phases of the speed-beat signals on the several frequencies. Various keyed Doppler systems[5] also provide varying amounts of range information on moving targets. Some of these systems remain operative in the presence of extremely strong unwanted signals from fixed targets. One keyed system, of possible interest but not yet developed, is that using a continuous signal with square-wave modulation or keying of its frequency. As mentioned in section 4i of Chapter II., this would give range information in terms of phase jumps in the Doppler-beat signal.

As a result of the development of pulse-radar techniques,

distance measurement by means of pulse receiver-retransmitters or transponders has come into use. Similar systems based on f-m radar techniques[6] are entirely possible. To measure the distance between two points in this way, a frequency-modulated transmitter and a receiver would be set up at one of them. At the second point, the frequency-modulated signal from the transmitter at the first point would be received and used to control another transmitter, also at the second location, which would retransmit, on the same or another frequency, a faithful replica of the received modulation. Received at the first point, this retransmitted frequency-modulated signal would be compared with that being transmitted directly. As in f-m radar, this comparison would yield a beat signal of frequency proportional to the time lag produced or distance covered in the round trip between the points in question. Where one transponder must cooperate in a number of independent distance determinations simultaneously, pulse technique seems more simply applicable than that of frequency modulation.

Another non-radar application of radar techniques is the determination of position in terms of the difference in time lag for propagation of signals over two distinct paths. In the pulse case, this application has become well known as the Loran system of navigation. An analogous use of f-m radar techniques has also been proposed.[7] In the f-m case, a single frequency-modulated radio signal would be simultaneously transmitted from two well separated reference points, with a controlled time delay in one transmission. The frequency of the beat developed between the two transmissions at a single receiving point would measure the difference of the distances from the two transmitting points to that receiving point.

The particular hyperbola, with the transmitting points as foci, which passes through the receiving point is determined in such a system by the difference of distances to the two transmitters. Intersection of two such hyperbolas, each determined by transmissions from a separate pair of reference points, determines completely the location of the receiving point (except for certain ambiguities). Rate of frequency change in modulation and delay of the later

transmission would be chosen, in relation to the separation of the transmitting points, to give convenient minimum and maximum values for received beat frequency. There would be no ambiguity as to which transmission was being received first if the delay applied to the later transmission exceeded that for propagation over the distance between the transmitting points. This is a system which might prove highly practical and useful if developed, because of the ease with which automatic flight can be controlled by using f-m radar techniques.

12. NOTATION AND REFERENCES

a. *Notation.* The algebraic notation listed alphabetically below has been used in this chapter.

a	Vertical acceleration (positive upward).
A	Aircraft altitude.
A_a	Value of altitude at the moment when it is measured.
A_m	Maximum altitude in operating range of equipment.
A_r	Value of altitude at moment bomb is released.
A_s	Value of altitude at moment slant range and speed are measured.
C	Capacitance in differentiating circuit.
d	Total difference in electrical length between feed through and mixing-signal paths from transmitter to mixer of f-m radar system.
D	Coefficient of slant speed in rocket-sighting approximation.
e	Total output voltage of radar counters, or of differentiator.
e_0	Bias component of radar output voltage.
e_1	Value of total radar output at which relay operation occurs; also, input voltage to differentiator.
e_2	Feed-back voltage in differentiating circuit.
f_m	Radar modulation frequency.
F_0	Mean radio frequency of radar transmission.
g	Downward acceleration of gravity.
G	Gain of feed-back amplifier in differentiating circuit.
h_R, h_s	Range and speed sensitivities of counters.
H_1, H_2	Limiting horizontal closing speeds of aircraft relative to target.
i	Current in differentiating circuit.

k	Ratio of frequency-sweep reduction in signal-following superheterodyne.
k_R, k_S	Range and speed sensitivities of radar.
K	Partial range coefficient in rocket-sighting approximation.
$L_1, \ldots, L_6$	Electrical-length components in feed-through and mixing-signal paths of radar system.
r	Resistance in differentiating circuit.
r_1, r_2	Voltage-dividing resistors in modulation-control circuit.
R	Slant range or distance between radar and target.
R_1	Range increment for special purposes.
R_r	Residual electrical range in transmission lines.
S	Slant speed of approach of radar to target.
S_o	Speed-axis intercept of linear range-speed relation.
t	Time.
T	Time interval required for radar to move from present position to vertical through target.
T_f	Time interval required for fall of bomb.
T_{min}, T_{max}	Operating time-to-target limits of equipment.
T_r	Time to target at instant of bomb release.
T_t	Change in time to target for release required to compensate bomb trail.
T'	Slope of linear range-speed approximation.
V	Vertical speed of aircraft (positive upward).
V_o	Vertical speed acquired in free fall from rest through altitude A.
V_r	Value of vertical speed at moment of bomb release.
V_s	Value of vertical speed at moment when slant range and speed are measured.
V_v	Value of vertical speed at moment when it is measured.
W	Width of radio-frequency band swept in frequency modulation.
W_{i-f}	Width of i-f band swept by residual modulation in signal-following superheterodyne.
W_{min}, W_{max}	Operating limits of width of modulation sweep.
α	Angular depression below horizon of line of sight from radar to target.
β	Angular depression below line of rocket launchers of line of sight to target.

β_o Reference value of rocket-sight depression angle.

τ Time constant of feed-back servo in differentiating circuit.

τ_a Time lag between occurrence of altitude value and availability of altitude datum for use.

τ_b Time lag in bomb-release mechanism.

τ_s Time lag between occurrence of slant-range and slant-speed values and occurrence of operation based on those values.

τ_v Time lag between occurrence of vertical-speed value and availability of vertical-speed datum for use.

ϕ Dive angle of aircraft path.

ψ Phase shift in i-f amplifier.

ω Angular frequency $2\pi f$.

b. *References.*

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CHAPTER IX.

MULTIPLE TARGET SYSTEMS

1. GENERAL CHARACTERISTICS

Frequency-modulated radar systems capable of distinguishing sharply among a multiplicity of targets may be used in either of two ways. They may be made to lock onto and track the motion of a chosen target without disturbance by other targets, or they may be used to search out and indicate the locations in space of a number of targets. Selective tracking methods were not studied extensively enough to merit discussion here, though reference may be made to the time tracker described in section 5 of Chapter VIII., which possesses some selectivity by virtue of its relatively narrow beat-frequency pass band.

Search systems of course invite comparison with the widely used pulse-radar systems. Searching of the space around the radar in various directions is accomplished by mechanical motion of a beam-forming radiating element, whether the radar is frequency modulated or pulse modulated. This causes the search beam to take up successively, or scan, all directions from the radar to those portions of the surrounding space that it is desired to investigate. Directional scanning in the case of f-m radar is distinguished from the pulse-radar case only by the increased importance of minimizing fortuitous coupling of f-m signal from transmitting to receiving channel, which may necessitate synchronized scanning with two antennas, and by the relatively infinitesimal amount of effort devoted to development for f-m use. Since direction scanning for f-m radar exhibits neither novelty nor distinction from pulse radar, it will not be discussed further.

Searching in range, for a single direction, is the function in which wide practical divergence among possible radar systems becomes evident. In pulse radar, range scanning results naturally and automatically from the very rapid inherent outward propagation of each radiated pulse

front. Indication follows equally naturally by moving an index, normally the point of impact of an electron beam on a fluorescent screen, along a convenient range scale. This index motion is in exact synchronism with the outward motion of the pulse front, but at a greatly reduced speed. At the instant that the pulse reflected from a target at a given range returns to the radar, it intensifies or displaces the moving index at the proper point of the scale to indicate presence and range of the target.

No such simple and natural method of range scanning and indication has become evident in the case of f-m radar. The methods used resemble rather those for "gating" pulsed radar. Each range corresponds to a different beat-note frequency, for any single condition of periodic frequency modulation. The required indicator is therefore of the nature of a spectrum analyzer, and as many f-m radar range-search systems can be contrived as there are possible types of analyzer.

Using a fixed condition of modulation, all range-beat frequencies may be applied simultaneously and continuously to a multiple-channel selective analyzer of fixed characteristics. This is the analogue of optical spectrum analysis by an ordinary prism or grating spectrograph. Indication may be by means of a multiple-element indicator, with one element continuously connected to each selector channel. One example of a combined multiple-channel selector and indicator is the well known Frahm reed frequency meter. Alternatively, each channel of the multiple selector may be connected in turn to a single-channel scanning indicator. Or, a single variable selector may be scanned across the beat-frequency output spectrum produced by the radar, in synchronism with the scanning of an indicator to show the selector output. This is the analogue of optical spectroscopy using a spectroscope as an adjustable monochromatic illuminator. It is also the analogue of range search with a movable gate in the case of pulse radar.

Another method useful for searching in range with f-m radar has no analogue either in optics or pulse technique. In this method a single fixed-frequency selector is used and the target spectrum is scanned past it by varying the range

sensitivity of the radar. Two modifications with somewhat different properties are possible, since the rate of change of transmitted frequency and correspondingly the radar range sensitivity may be varied by adjustment either of the frequency band swept in modulation or of the modulation frequency. Simpler equipment than is necessary for the multiple-channel method suffices for a single-channel selector with a range-sensitivity scan, but performance is better in the multiple-channel case.

Table IX.-1 exhibits a number of ways in which f-m radar search systems might be arranged. Some of these systems will later be compared in theory as to signal/noise ratio, limiting range, range resolution,

TABLE IX.-1

Classification of F-M Search Systems

System Class	*Modulation Sweep*	*Modulation Period*	*Filter Type*	*Filter Tuning*	*Filter Band*	*Indicator Type*	*Range Resolution*
A_1	Fixed	Fixed	Multiple	Fixed	Fixed	Multiple	Arbitrary
A_2	Fixed	Fixed	Multiple	Fixed	Fixed	Scanning	Arbitrary
B_1	Fixed	Fixed	Single	Variable	Fixed	Scanning	Constant Increment
B_2	Fixed	Fixed	Single	Variable (heterodyne)	Fixed	Scanning	Constant Increment
B_3	Fixed	Fixed	Single	Variable	Variable	Scanning	Arbitrary
C_1	Variable	Fixed	Single	Fixed	Fixed	Scanning	Constant Fraction
C_2	Variable	Fixed	Single	Fixed	Variable	Scanning	Arbitrary
C_3	Fixed	Variable	Single	Fixed	Fixed	Scanning	Constant Fraction
C_4	Fixed	Variable	Single	Fixed	Variable	Scanning	Arbitrary

and rapidity of action; effects of target motion will be neglected in such comparisons. One system was built in experimental form as the *AN/APQ-19* equipment and was given limited field tests; this equipment, of the class designated C_1 in the Table, will be further described. Methods of avoiding confusion by target motion and of distinguishing moving targets from stationary ones will also be discussed.

One general characteristic of target-selective f-m radar requires mention before special cases are discussed. Periodicity of the radar modulation requires that the selectors be repeatedly excited at frequent intervals. Selectors inherently require a finite time to respond to new excitation, and the delay so occasioned is proportional to the sharpness of selection. Each frequency sweep must last long enough to permit the selectors to develop an adequate and definite response. There is thus a fundamental limitation to the rapidity with which frequency-modulated radar can gather information. The analogous limitation that occurs in the case of pulse radar proves less restrictive under practical conditions.

2. Performance and Parameters of Search Systems

a. *General.* For purposes of later comparison, some aspects of the performance of pulsed and variously frequency-modulated search radars will now be calculated. A number of assumptions must be made concerning each of the systems to be compared, and the results obtained[1,2] are considerably influenced by the details of these assumptions. The final comparisons must therefore be regarded as only qualitatively significant.

Effects of directional scanning should not depend upon type of modulation, so comparisons will be made only for the case of range scanning with a fixed directional beam, or "searchlighting". Directional scanning would reduce maximum working range for any of the systems considered, probably to a similar degree for each. Effects of directive-antenna power gain and of echoing area of target likewise should not depend upon type of modulation or indication, so will be included as a typical factor common to all cases.

For the 18-inch parabolic reflectors designed for the

AN/APS-3 pulse radar and used also in the *AN/APQ-19* f-m search system, the effective antenna area A_a' is 20 square wave lengths (geometrical area $29.2\lambda^2$) at the 7.5-cm. wave-length used in the *AN/APQ-19*. For a medium-sized surface vessel acting as a target at that wave length, a mean effective scattering area of 8000 square meters may be assumed. With such antennas used against such a target, equation (II.33) indicates that an available echo power P_r of $1.21 \times 10^{-10} R^{-4}$ watts will be received, at a range of R nautical miles, per watt radiated by the transmitting antenna (one nautical mile is 6080 feet).

Comparisons will be made on a basis of equal range resolution, so far as different types of resolution obtained with different systems permit. Data-gathering time intervals will also be held equal so far as inherent differences permit. An average radiated power of one watt will be assumed in each case for purposes of comparison.

b. *Pulse Case.* To make the field of comparison complete, some pulse-radar performance factors will first be set up. Transmitted pulses having rectangular form and lasting one microsecond will be assumed. These will be converted by a receiver having an optimum selectivity characteristic to triangular output pulses 2 microseconds wide at the base, so that targets separated in range by 1000 feet or more will be definitely resolved (that is, indicated as distinctly separate). The indefiniteness of meaning of some of the assumptions here made is well illustrated by the fact that this pulse may resolve on occasion targets separated by as little as 500 feet, though its typical resolving power will be considered to be 1000 feet. The total transmitted energy per pulse, for the shape and duration assumed, is $10^{-6}P_p$ joules, where P_p is r-m-s power in watts at the peak of the transmitted pulse envelope. For an average transmitted power of one watt and a repetition frequency of f_r pulses per second, the energy per pulse will be $1/f_r$ joules and peak power will be $10^6/f_r$ watts.

Extensive empirical analysis of actual A-scope pulse observations[3] has indicated that, in order to be just discernible in the presence of random noise, each rectangular received pulse must have a total energy of at least U_{min},

where

$$U_{min} = U_N (1670/f_r)^{1/3} \left[\sqrt{t_p \Delta f} + 1/\sqrt{t_p \Delta f}\right]^2 /4. \qquad \text{(IX.1)}$$

U_N is the equivalent noise power per unit frequency band at the receiver input, as determined from the receiver-output noise power in terms of the gain of the receiver and its noise band width Δf; t_r is the duration of the rectangular input pulse.

An unpublished theoretical study by D. O. North has led to the similar but more general conclusion that

$$U_{min}/U_N = S\sqrt{D}/(K\sqrt{n}) + \epsilon . \qquad \text{(IX.2)}$$

Here, S is a factor depending on the statistical criterion chosen for a just-discernible signal, D is a factor characteristic of the type of detector used and never in practice far from unity, and ϵ is a term occurring for visual observation and related to the finite contrast threshold of the eye. K is a function of which the form depends on the shape of the pulses and that of the receiver-selectivity characteristic; for definite shapes, the value of K depends only on that of the product $t_p \Delta f$ (for other than rectangular pulses, t_p may be defined as duration of a rectangular pulse having equivalent energy and peak power). Since too broad a receiver will pass excessive noise without increasing the signal, while too sharp a receiver will reject some of the signal without comparably reducing noise, K will always exhibit a maximum for some optimum value of $t_p \Delta f$ near unity. With receiver noise band Δf designed to be just $1/t_p$, the factor K in the empirical form $4/\left[\sqrt{t_p \Delta f} + 1/\sqrt{t_p \Delta f}\right]^2$ reaches its maximum value of unity.

The number n of pulses integrated by the eye and brain, or other observing mechanism, is $T_0 f_r$ if T_0 is the effective time of integration. Integration of coherent data from many pulses of course reduces the disturbing effect of random noise. Because of the additive term ϵ, the inverse-square-root dependence of U_{min}/U_N on f_r or n given by the theory may be quite compatible over a limited region with the observed apparent dependence on inverse cube root of f_r. Other observations[4] support within limits the inverse-square-root form. The difference is of little practical importance, and the better known $f_r^{-1/3}$ form of (IX.1)

will be used in the comparisons made here.

For a noise temperature T of 300°K. and an assumed receiver noise figure of 40 (16 db. above thermal, for a receiver which does not reject image response), the noise power U_N per unit band will be $40kT$ or $1.64\text{x}10^{-19}$ joules. In the case under consideration, the received pulse energy is $1.21\text{x}10^{-10}R^{-4}f_r^{-1}$ joules, and by (IX.1) must at least equal $1.64\ (1.67)^{1/3}\text{x}10^{-18}f_r^{-1/3}$ joules in order to be observed in the presence of random noise. The range difference δR resolved by this pulse radar is 1000 feet, all data is repeated at time intervals $1/f_r$, and noise-limited operation is possible out to a maximum range R_{max} of $88.8f_r^{-1/6}$ nautical miles. At a normal radar-pulse repetition frequency of 2000 per second, peak pulse power will be 500 watts and a maximum range of 25 miles is to be expected. At a low repetition frequency of 10 per second, peak pulse power must be 100 kilowatts and the corresponding maximum range is 60.5 nautical miles. An intermediate-frequency noise-pass band Δf of one megacycle is required in the receiver, to satisfy the optimum condition in $t_p\Delta f$ for the postulated 1-microsecond rectangular received pulse.

Both total energy per pulse and average transmitted power must be as large as possible to maximize the noise-limited range of a pulse radar. For a fixed available pulse energy, a high repetition frequency is desirable to assure adequate average power; this fact is quite well known. In the present discussion available average power is held fixed, with the consequence that a low repetition frequency is found desirable to assure adequate pulse energy; this aspect of the matter is not so well known. The practical difficulty of supplying sufficient energy per short pulse to maintain reasonable average power at very low repetition frequencies is disregarded here. With receiver frequency-response characteristic kept properly matched to pulse duration, maximum range is determined independently of range resolution in the case of pulse radar.

c. ***Multiple-Channel F-M Case.*** It was shown in section 41 of Chapter II. that the nature of the frequency modulation used sets a fundamental limit to the minimum

separation at which targets can be individually distinguished by f-m radar. This limiting separation δR_t analogous to the fixed error of single-target systems, is of the order of $\frac{1}{2}c/W$ for a frequency-modulation band of width W; this is one half the *sweep wave length*, or wave length of a signal of frequency W. A range difference of $\frac{1}{2}c/W$ will, according to the basic range-sensitivity equation (II.22), produce a range-beat frequency difference of $2f_m$, with triangular frequency modulation at a frequency f_m. Selection must start afresh with each half cycle of modulation, and a selective circuit sharp enough to separate frequencies differing by $2f_m$ can only come into full operation after a time lag of the order of $1/(2f_m)$. That is, the fundamental resolution limit corresponds in practice to the selection capability of a filter which has only a limited time to act.

In order of magnitude, a selective circuit having a *noise band width* Δf will reach a substantially steady operating state in a time $1/\Delta f$ after a steady signal is applied to it (overall time lag, which delays but does not distort the signal, is here neglected). The exact relationship will depend on circuit details. It will be assumed here for purposes of comparison that a selective circuit with response time $1/\Delta f$ will distinguish definitely between signals differing in frequency by $\frac{3}{2}\Delta f$; this is at least the right order of magnitude for the resolution attained. Complicating effects of Doppler frequency change caused by target motion will be neglected for the present.

Direct empirical data for evaluation of minimum discernible signal is totally lacking in the case of f-m radar. Analogy with pulse operation permits some conclusions to be reached, however. For a single, isolated, linear sweep of transmitter frequency, the received signal from any one target at fixed range gives rise when mixed with the transmitter output to a burst or pulse of beat-note signal of constant frequency. This pulse of signal, at the range frequency for that target, has a rectangular envelope. Periodic frequency modulation produces a series of such beat-note pulses, at a repetition frequency f_r which is $2f_m$ for triangular or symmetrical-sawtooth modulation.

Pulse duration is $1/(2f_m)$, reduced by time lost at modulation turn around. Lost time τ required for signal propagation (no other time loss will be considered here) is $2R/c$, and may be specified as a fraction ρ of the duration of a single sweep. The useful pulse duration t_p is then $(1-\rho)/(2f_m)$, or $(1-\rho)/f_r$. These long beat-note pulses, following each other without other pause than the time lost at turn around, will be treated in just the same way as ordinary radar pulses. Each of the multiple selective channels may be treated as a pulse-radar receiver, when a single indicator scanning all channels once per single modulation sweep is used. This is a system of class A_2 of table IX.-1. A multiple-channel indicator might yield slightly different results.

According to (IX.1), the noise band of each channel should be $2f_m/(1-\rho)$ cycles per second, for optimum response ($t_p\Delta f=1$) to pulses lasting $(1-\rho)/(2f_m)$ second. The minimum beat-frequency difference for target resolution ($3/2\Delta f$) is then $3f_m/(1-\rho)$. By equation (II.22), this corresponds to a range resolution δR of $3c/[4(1-\rho)W]$ or $\frac{3}{4}\lambda_w/(1-\rho)$, which is $737/[(1-\rho)W]$ feet if W is in megacycles per second, Modulation sweep width W alone thus determines the range resolution, so long as $\rho \ll 1$. Under the same condition, modulation frequency f_m alone determines the pulse duration t_p, hence with a constant transmitted power level P_t the total transmitted energy $P_t t_p$ per pulse, as well as the data-integration factor $(1670/f_r)^{1/3}$. f_m thus sets the signal/noise ratio of the system for a given range. Resolution and maximum range are therefore separately controllable in multiple-indicator f-m radar, as in the case of ordinary pulse radar.

Beat-frequency pulses are applied to all filters in quick succession, without major idle periods. Decay of the filter response to the preceding pulse therefore affects its response to each currently occurring pulse. Since successive beat pulses are substantially incoherent in phase unless range is almost perfectly constant, this effect of previous-pulse decay is normally a random one and amounts to an increase in noise level. The added noise caused by previous pulses is small if the filters reach substantially a steady state during each pulse.

For a resolution of 1000 feet, a sweep width W of $0.737/(1-\rho)$ megacycle is required. All data is repeated at time intervals of $1/(2f_m)$ or $1/f_r$. Received energy per pulse is $1.21\times10^{-10}R^{-4}(1-\rho)f_r^{-1}$ joules for a steady transmitted power of one watt. Minimum discernible pulse energy is $1.95\times10^{-18}f_r^{-1/3}$ joules for a receiver (with no image rejection) having a noise figure of 40 and a noise temperature of 300°K. If ρ can be neglected, equating received energy to minimum discernible energy again gives just the maximum range $88.8f_r^{-1/6}$ miles found in the pulse-radar case.

Results that do not neglect the lost-time fraction ρ can be obtained quite easily. Equation (IX.1) for minimum discernible beat-note pulse energy is still directly usable (with $t_p\Delta f$ still unity), but received-pulse energy as determined from equation (II.33) should now be expressed in terms of ρ. Since ρ is by definition $2f_rR/c$, this process gives

$$U_r = t_pP_r = \frac{P_tA_a'^2A_e'}{2\pi F_0}\cdot\left(\frac{2f_r}{F_0}\right)^3\cdot\frac{1-\rho}{\rho^3} \qquad \text{(IX.3)}$$

as the received energy per beat-note pulse. Repetition frequency f_r is found in terms of lost time ρ_m at maximum noise-limited range, by equating energies as given by (IX.1) and (IX.3), to be

$$f_r = f_1\left\{\frac{P_tA_a'^2A_e'}{2\pi F_0\overline{NF}kT}\right\}^{-3/10}(2f_1/F_0)^{-9/10}\rho_m^{9/10}(1/\rho_m - 1)^{-3/10}. \qquad \text{(IX.4)}$$

Maximum range is then

$$R_{max} = \frac{\rho_m c}{2f_r} = \lambda_0\left\{\frac{P_tA_a'^2A_e'}{2\pi F_0\overline{NF}kT}\right\}^{3/10}(2f_1/F_0)^{-1/10}\rho_m^{1/10}(1/\rho - 1)^{3/10}. \qquad \text{(IX.5)}$$

F_0 is mean radio frequency and λ_0 the corresponding wave length, while f_1 is 1670 cycles per second, the empirical reference value of repetition frequency in equation (IX.1).

Assuming again a transmitted power of one watt, radio frequency of 4000 megacycles per second, noise figure of 40, noise temperature of 300°K., effective antenna area of 20 square wave lengths, and effective target area of 1.42×10^6 square wave lengths (λ_0 of 7.5 cm.), the dimensionless coefficient within the curly brackets of each of

the above equations has the value 1.38x10[17]. Inserting numerical values,

$$f_r = 3560\,\rho_m^{9/10}(1/\rho_m - 1)^{-3/10} \text{ cycles per second} \qquad \text{(IX.4a)}$$

and

$$R_{max} = 22.7\,\rho_m^{1/10}\,(1/\rho_m - 1)^{3/10} \text{ nautical miles.} \qquad \text{(IX.5a)}$$

These equations express the implicit relation between R_{max} and f_r in conveniently calculable parametric form. The explicit relation between R_{max} and f_r directly is a cubic equation and so is not convenient for calculation.

At a repetition frequency of 2000 per second, or modulation frequency of 1000 cycles, (IX.4a) indicates that 51.5 per cent of the time is lost in propagation. This loss reduces maximum range from 25 to 21 nautical miles, and increases the modulation sweep required for 1000-foot resolution from 0.74 to 1.52 megacycles per second. At 10 repetitions or 5 cycles of modulation per second, the lost time is only ¾ per cent, reducing maximum range by a negligible amount from the pulse-radar value of 60.5 miles, and increasing required sweep only from 0.74 to 0.75 megacycles.

Let R_{max} be the mean range of the maximum-range channel, which responds to targets at ranges within a region of depth δR. Likewise, let R_{min} be the mean range of the minimum-range channel, and let all filters have equal pass bands, so that each one of the multiple channels will have the same range resolution δR. The number M of filters or range channels required will then be simply

$$M = 1 + (R_{max} - R_{min})/\delta R\,. \qquad \text{(IX.6)}$$

Individual channels should be tuned to frequencies differing by $\frac{3}{2}\Delta f$, which is $3f_m/(1-\rho)$. With 1000-cycle modulation (2000 repetitions per second), 127 channels each 1000 feet wide would be required to give operation from a minimum-range channel centered at 500 feet out to a maximum range of 21 nautical miles; 100 channels would permit operation down to a minimum range of 4.7 miles. With 10 repetitions per second (5-cycle modulation), 367 channels would give operation from zero range (500 feet) out to 60.5 nautical miles.

The quantities f_r, ρ_m, and R_{max} are all closely related

for noise-limited "searchlighting" operation of a multiple-channel f-m search radar. If any one of these characteristic quantities is assigned a definite value, the other two are definitely determined. The pair of quantities W and δR are also closely related, so that if one of them is now assigned a value (ρ having already been specified), the other is determined. The quantities M and R_{min} are related, and specification of either (R_{max} being already determined) fixes the other. That is, if one characteristic in each of the above three groups is specified, the limiting performance of the system is entirely determined. The performance parameters R_{min}, R_{max}, δR and f_r are likely to be directly specified and the apparatus characteristics ρ_m, M, W, and transmitted power to be determined indirectly by such specification. If an unreasonable value of some apparatus characteristic results, the performance specification must be modified.

d. *Single Channel with Variable Tuning.* Scanning in range with a frequency-modulated radar having only one selective channel introduces new limitations and new relationships. Perhaps the simplest single-channel case is that in which modulation conditions are fixed and tuning of the selective channel is varied. This may be done for example by observing the beat-note output of the radar oscilloscopically with the aid of a wave analyzer of the well known heterodyne type, which passes a frequency band of constant width. The oscilloscope is to be swept in synchronism with the wave-analyzer tuning, so as to produce the widely used "A-scope" type of radar indication, with target strength plotted on one axis against range on the other axis of the oscilloscope screen. This method of range searching belongs to class B_2 of table IX.-1 and is analogous to use of a sweeping range gate in pulse radar.

Each successive increment of range to be resolved will be examined separately for the duration of one single sweep of transmitted frequency. Repetition frequency f_r of the individual sweep must now be distinguished from the frequency f_s at which the complete range search or scan is repeated. In the pulse-radar and multiple-channel f-m radar systems, with all range elements searched during each single sweep, there was no distinction between search-

repetition frequency f_s and pulse-repetition frequency f_r. In the single-channel f-m radar scanning in range, f_r or $2f_m$ must evidently equal Nf_s, where N is the number of range elements to be searched in succession. N is now given in terms of R_{max}, R_{min}, and δR by equation (IX.6), just as M was in the multiple-channel case.

Useful duration of the received beat-note pulses, which determines both the useful energy transmitted per pulse for a fixed average power and the usable selectivity of the single filter, is again $(1-\rho)/f_r$ or $(1-\rho)/(2f_m)$. Equation (IX.3) therefore still holds for received-pulse energy. The frequency which controls data integration is now that of the complete search, however, so f_s must replace f_r in using equation (IX.1) to fix the minimum discernible signal.

If lost-time fraction ρ is negligible compared to unity, setting received-pulse energy equal to minimum-discernible energy again gives simple power-law expressions for R_{max}. With the usual values for power, wave length, antenna and target areas, and receiver noise, R_{max} is $88.8f_s^{1/12}f_r^{-1/4}$, $88.8N^{1/12}f_r^{-1/6}$, or $88.8N^{-1/4}f_s^{-1/6}$ nautical miles. The performance penalty paid for repeated use of a single selective channel is particularly evident from the last of these expressions, which with $N=1$ applies also to pulse-radar systems and multiple-channel f-m systems. Successive scanning of 81 range elements reduces maximum noise-limited range to one third of its multiple-channel value, other things being equal.

Holding f_s constant, a convenient explicit relation between f_r and R_{max} directly can easily be obtained from (IX.3) and (IX.1), without neglecting ρ. Since f_r is not likely to be arbitrarily chosen in practice, however, such a relation is not very useful. Practically important apparatus-design variables are f_s, which determines the speed of data gathering, and N, which determines the number of points to which the single filter must be tunable. Equating energies from (IX.1) and (IX.3), with f_r replaced by Nf_s in the latter, there result instead of (IX.4a) and (IX.5a)

$$f_s = 3560N^{-9/10}\rho_m^{9/10}(1/\rho_m - 1)^{-3/10} \text{ cycles/second} \quad \text{(IX.7)}$$

and

$$R_{max} = 22.7N^{-1/10}\rho_m^{1/10}(1/\rho_m - 1)^{3/10} \text{ nautical miles.} \quad \text{(IX.8)}$$

For multiple-channel operation, with only one modulation sweep per complete search, these expressions reduce immediately to (IX.4a) and (IX.5a)

One quantity alone no longer suffices to determine ρ_m and R_{max}, but any two of the three simply related quantities f_s, N, and f_r will determine the third, and ρ_m and R_{max} as well. Once N and f_s, for example, have been specified and ρ_m and R_{max} have been determined from them, one and only one of the further related quantities δR, W, or R_{min} may also be specified. The noise-limited performance of the system is then completely determined. The final choice is not entirely free, since a value of W or δR that made R_{min} negative would not represent proper use of the N range elements provided by the radar. Nor may N and f_s be chosen so large that (IX.7) gives an excessive value for lost-time fraction ρ_m at maximum range. Excessive lost time would require an excessive width W of frequency sweep to realize a given range resolution δR, since

$$W = 3c/[4(1-\rho)\,\delta R]\,. \tag{IX.9}$$

Specification of performance parameters R_{min} and δR would be preferable to specification of N, but would not lead so directly to a complete design.

If 100 range elements are searched in succession 10 times per second (modulation frequency 500 cycles per second), equation (IX.7) indicates that 22 per cent of the time is lost in propagation at maximum range. Equation (IX.8) then gives 17.8 nautical miles as the maximum range on the 8000-square-meter target used in these comparisons. From (IX.9), the frequency sweep required for 1000-foot resolution is 0.95 megacycles per second. R_{min}, which is $R_{max}-(N-1)\delta R$, is then 1.55 nautical miles or 9400 feet. The range-beat frequency spectrum to be analyzed extends from 18 to 208 kilocycles per second, the required noise band of the filter being 1280 cycles per second.

Resolution at less than maximum range could be improved in the single-channel, variable-filter system by changing the filter selectivity to agree with the value of ρ for each range element. Different selectivities might be employed similarly in the separate channels of a multiple-channel system. This refinement would be a considerable complica-

tion and would not improve performance at maximum range, so will not be considered further.

e. *Single Channel with Variable Sweep Width.* It is especially convenient in practice to use a single fixed-tuned selective channel and to scan the range spectrum over the pass band of this filter, by control of the rate of change of transmitter frequency in modulation. Rate of change of frequency is conveniently controlled by varying the width of frequency-modulation sweep, while maintaining constant duration of sweep (that is, constant modulation frequency). This is the system classed as C_1 in Table IX.-1. Rate of change of frequency should alter suddenly between successive linear sweeps, with accurately constant rate during each sweep.

Useful pulse duration t_p at maximum range will again be $(1-\rho_m)/f_r$, where f_r is again $2f_m$, and is matched by a filter noise band Δf of $1/t_p$ cycles; resolved frequency difference will again be taken as constant at $3f_m/(1-\rho_m)$. The resolved range difference δR will therefore be again $\frac{3}{4}\lambda_w/(1-\rho_m)$, while actual range R observed with a filter pass band centered at a fixed frequency f_0 will be $cf_0/(4Wf_m)$. Since ρ is $4f_mR/c$, f_0 must be just ρW. Range resolution δR now varies as W is changed but the fractional resolution $\delta R/R$, which may be called simply δ, is given by

$$\delta = \delta R/R = 3f_m/[(1-\rho_m)f_o] \tag{IX.10}$$

and is obviously constant throughout the range search. From one range element to the next, range increases by the constant factor $1+\delta$. N is again the total number of resolvable range channels scanned in succession, and the limiting channels are again centered on R_{min} and R_{max}.

If one range element is observed for each successive single sweep, if range search proceeds outward, and if time is measured from the start of each complete range scan, then

$$R_i = R_{min}(1+\delta)^{2f_m t_i}. \tag{IX.11}$$

R_i is the central range of the i^{th} region scanned and t_i is the time at the start of the i^{th} frequency sweep. The first sweep is that centered on R_{min} and the N^{th} that centered on R_{max}; t_i is always an integral multiple of

$1/(2f_m)$, the multiplying integer being $i-1$. Width of the modulation sweep has its largest value W_{max} at minimum range, and sweep width at R_i is

$$W_i = W_{max}(1+\delta)^{-2f_m t_i} \quad . \tag{IX.12}$$

Since only those values of t which are integer multiples of $1/(2f_m)$ are used, this represents a stepped exponential variation of W with time. The search repetition frequency f_s is again $2f_m/N$, or f_r/N.

Since fractional range resolution $\delta R/R$ rather than incremental resolution δR remains constant throughout the search in the variable-sweep-width system with fixed filter selectivity, any comparison with systems of constant incremental resolution must be somewhat arbitrary. The comparison among systems will be made here on a basis of equal δR at arithmetic mean range. In this case

$$\delta = 2\,\delta R_{mean}/(R_{max}+R_{min}) \ , \tag{IX.13}$$

while for constant fractional increment δ,

$$R_{max}/R_{min} = (1+\delta)^{N-1} \quad . \tag{IX.14}$$

Both δ and R_{min} can be determined by simultaneous solution of (IX.13) and (IX.14). Choice of geometric mean range for comparison would have led to undue complication, as would comparison at minimum range. Choice of either minimum or maximum range for comparison would be unfair to one or the other system.

Equations (IX.1) (with f_s in place of f_r) for minimum discernible pulse energy, and (IX.3) for received beat-note pulse energy, both remain valid for the single-channel system scanned in range by variation of sweep width at constant modulation frequency. Equations (IX.7) and (IX.8) may therefore be used again to determine R_{max} and ρ_m, for given values of search-repetition frequency f_s and number N of range elements successively scanned. For 100 elements scanned 10 times per second (500-cycle modulation), ρ_m is again 22 per cent and R_{max} is 17.8 nautical miles.

With δR at mean range 1000 feet, $\delta R_{mean}/R_{max}$ is 0.00922, for which (IX.13) and (IX.14) indicate a fractional resolution δ of 0.0150, or 1.5 per cent, and a value of 4.37 for

the ratio of R_{max} to R_{min}. Minimum range under the stated conditions of operation is therefore 4.1 miles. Incremental resolution δR is 370 feet at minimum range and 1630 feet at maximum range. Using (IX.9), the modulation sweep required for this resolution is 0.58 megacycle per second at maximum range and 2.54 megacycles at minimum range. The frequency f_0 to which the fixed filter is tuned must be ρW or 128 kilocycles, with a pass band Δf of 1.28 kilocycles per second.

It should be evident that a system giving constant fractional resolution is undesirable when operation down to very small minimum range is required. Because of the large ratio of R_{max} to R_{min} required, either poor fractional resolution or successive examination of a very large number of elements per scan is necessary for use of such a system to extreme short range. Successive scanning of many elements results either in reduction of maximum noise-limited range or in slow gathering of data, or both.

f. *Single Channel with Variable Sweep Duration.* In a search system using a single selector tuned to a fixed beat frequency f_o, rate of change of transmitted frequency during modulation may alternatively be controlled by varying the time interval during which that frequency is swept uniformly across a band of fixed width W. This interval, corresponding to $1/(2f_m)$ or $1/f_r$ in the case of fixed-frequency triangular-wave modulation, may be called t_r. t_r will be short for short ranges and long for long ranges, so the pass band of the selector should not be left fixed during search. If the filter were made sharp to match long pulses of beat signal and then left alone, for best signal/noise ratio at maximum range, it would not have sufficient time to respond well to the short pulses at minimum range.

The single-sweep duration t_r producing filter-center beat frequency f_o for range R is $2WR/(cf_o)$ and propagation time τ is $2R/c$, so

$$\rho = \tau/t_r = f_o/W \, . \qquad \text{(IX.15)}$$

Since both f_o and W are fixed, ρ is obviously constant as t_r is varied to search in range; ρ will therefore be written without subscript hereafter. If the filter noise band Δf is matched afresh to the duration of useful beat

signal for each individual modulation sweep, Δf will always be $1/[(1-\rho)t_r]$. If frequency difference for adjacent targets received by the filter is again $\frac{3}{2}\Delta f$, resolved range δR will again satisfy equation (IX.9) and will not vary during range scan. If any two of the quantities δR, f_0, W, and ρ are specified, (IX.9) and (IX.15) determine the other two.

Number N of successively resolved range elements is again given by (IX.6). Mid-channel range R_i observed during the i^{th} sweep of an outward range scan is

$$R_i = R_{min} + (i-1)\delta R = R_{max} - (N-1)\delta R + (i-1)\delta R \ . \qquad \text{(IX.16)}$$

Converting range to sweep time by the factor $2W/(cf_0)$ and using (IX.15), duration of the i^{th} sweep is

$$t_i = 2(R_{max} - N\delta R + i\delta R)/(c\rho) \ . \qquad \text{(IX.17)}$$

For the N^{th} sweep, at maximum range, t_N becomes simply $2R_{max}/(c\rho)$.

Time T_i from the start of a range scan to the start of the i^{th} modulation sweep of that scan is

$$T_i = \sum_{1}^{i-1} t_j = 2(R_{max} - N\delta R + \tfrac{1}{2}\, i\delta R)(i-1)/(c\rho) \ . \qquad \text{(IX.18)}$$

By elimination of i from (IX.17) and (IX.18) and solution of the resulting quadratic, the required parabolic variation of sweep duration with time of occurrence of the sweep in the scan can be found if desired. Total duration T_s of scan is given as T_{N+1} by (IX.18) and is

$$T_s = 1/f_s = 2N[R_{max} - (N-1)\delta R/2]/(c\rho) \ , \qquad \text{(IX.19)}$$

or

$$T_s = N(R_{max} + R_{min})/(\rho c) \ . \qquad \text{(IX.19a)}$$

This is just N times the arithmetic mean of the limiting sweep times t_1 and t_N.

Equation (IX.1), using f_s, still governs minimum discernible beat-note pulse energy. Equation (IX.3), with f_r replaced by $1/t_r$, also remains valid for received beat-note pulse energy, but should in this case be written as

$$U_r = \frac{P_t A_s'^2 A_s'}{2\pi F_0} \left(\frac{\lambda_0}{R}\right)^2 (1/\rho - 1) \ . \qquad \text{(IX.20)}$$

Setting received and minimum-discernible energies equal, there now results

$$R_{max} = \lambda_o \left\{ \frac{P_t A_a'^2 A_e'}{2\pi F_0 \overline{NF} kT} \right\}^{1/3} (f_s/f_1)^{1/6} \left({}^1/_\rho - 1 \right)^{1/3} \qquad \text{(IX.21)}$$

rather than (IX.5). With the usual comparison values of power, area, wave length, receiver noise, and data-integration reference frequency f_1, this is

$$R_{max} = 9.16 f_s^{1/6} \left({}^1/_\rho - 1 \right)^{1/3} \qquad \text{(IX.21a)}$$

Using the relation (IX.19) between f_s and R_{max} in conjunction with (IX.21) gives in turn

$$N^2 - 2N \left\{ \left[\frac{P_t A_a'^2 A_e'}{2\pi F_0 \overline{NF} kT} \right]^{1/3} (\lambda_o/\delta R)(f_s/f_1)^{1/6} \left({}^1/_\rho - 1 \right)^{1/3} + \frac{1}{2} \right\}$$

$$+ (\lambda_o/\delta R)(\rho F_o/f_s) = 0 . \qquad \text{(IX.22)}$$

Because of the complicated relation between t_N and f_s when searching in range by variation of sweep duration, the simplicity of equation (IX.4) or (IX.7) is lost. N and f_s alone no longer suffice to determine ρ and R_{max}. δR must be specified in addition to N and f_s in order to make such determination from the rather inconvenient simultaneous equations (IX.22) and (IX.21). Once all three characteristics are specified and ρ is thereby determined, the constant modulation sweep width W and filter frequency f_0 can be found from (IX.9) and (IX.15).

For operation with 100 range elements each 1000 feet deep and a search-repetition frequency of 10 per second, (IX.22) indicates a lost time of 15.7 per cent. Maximum range, from (IX.21), is then 20.8 nautical miles, with a minimum range of 4.5 miles which is less by $(N-1)\delta R$. W is 0.875 megacycle per second, and equation (IX.17) gives the sweep duration required as 1/2820 second at minimum and 1/610 second at maximum range. Filter pass band Δf is correspondingly 3350 to 715 cycles per second, with a fixed center frequency f_0 of 137 kilocycles per second. If δR remained 1000 feet and R_{min} were made 500 feet, by

choice of N and f_s, targets would in principle be observable down to zero range. Operation to such a low minimum range would require a large ratio of maximum to minimum sweep duration, as well as of maximum to minimum filter pass band, and might prove rather difficult to attain.

3. COMPARISON OF SYSTEMS

a. *Noise Limited Range.* Expressions for maximum noise-limited range for five different search-radar systems were developed in the preceding section. Values for maximum range were determined from these expressions for the five systems under the conditions of operation given in Table IX.-2. So many assumptions had to be made, however, that the absolute values found are significant only in order of magnitude.

TABLE IX.-2

Operating Conditions for Range Comparison

Frequency	4000 megacycles per second
Average Transmitted Power	1.0 watt
Antenna Power Gain	250 (effective area $20\lambda^2$)
Target Echoing Area	8000 square meters
Receiver Noise Figure	40 (16 db.), with no image rejection
Selectivity	Set for best signal/noise ratio at maximum range
Range Difference Resolved	1000 feet (at mean range if not constant)
Frequency of Complete Search	10 per second
Number of Range Elements	100

Pulse radar and frequency-modulated radar using multiple selective channels are both found to permit rapid search and to give substantially the same maximum range. The pulse system is slightly the better when an appreciable part of the repetition period is required for signal propagation. Range for these two systems is respectively 25 and 21 nautical miles when making 2000 complete range-search scans per second, or 60.5 miles for either at 10 scans per second. Three f-m radar systems using a single selective channel scanned in range by different methods

all permit operation only at very low scan frequencies and all show similar maximum ranges. Search by variable duration of frequency sweep permits somewhat more rapid operation or greater range than does either of the other two single-channel methods considered. Range with the three single-channel systems under the conditions considered is respectively 18, 18, and 21 nautical miles at 10 complete search scans per second; this is one third of the range obtained with the pulse-radar and multiple-channel f-m systems at the same low scan frequency.

All five systems require approximately a one-megacycle radio-frequency channel width to produce the specified range resolution, with the f-m systems penalized in channel width for time lost in propagation. R-f channel width is also the noise band width of the pulse system. Noise band width of the frequency-modulated systems is substantially independent of radio- or intermediate-frequency channel width, because no real detection takes place in such systems. There is at the "detectors" only frequency changing of weak incoming signals and noise, in the presence of strong local mixing signals. F-m noise band is that of the radar-beat-frequency channel only and is very narrow, being $2f_m/(1-\rho)$. For the modulation frequencies usable in practical f-m systems, this noise band may be from a few cycles to a few thousands of cycles per second wide, depending on the required resolution, number of range elements, and speed of operation.

High peak power in pulse-radar systems on the one hand and narrow noise band in frequency-modulated radar systems on the other hand give the two types of modulation comparable performance as to maximum noise-limited range. Inefficient use of the noise band by desired signal in the simple single-selector f-m systems puts them at a decided disadvantage in comparisons of performance with either pulse or multiple-selector f-m systems.

The various relations found in the preceding section to govern noise-limited range lend themselves well to graphical comparison. For the conditions of Table IX.-2, Fig. IX.-1 makes this comparison. Plotted to logarithmic scales of both maximum range and search-repetition frequency, the simple inverse-sixth-root relation characteristic of pulse

radar when operated at constant average power appears as a straight line in the figure. Frequency-modulated radar that performs a complete range search on each sweep of transmitted frequency shows essentially the same linear characteristic at low repetition frequencies. At higher search-repetition frequencies, however, loss of useful beat signal during radio-signal propagation causes maximum range for f-m radar to fall below that for pulse radar.

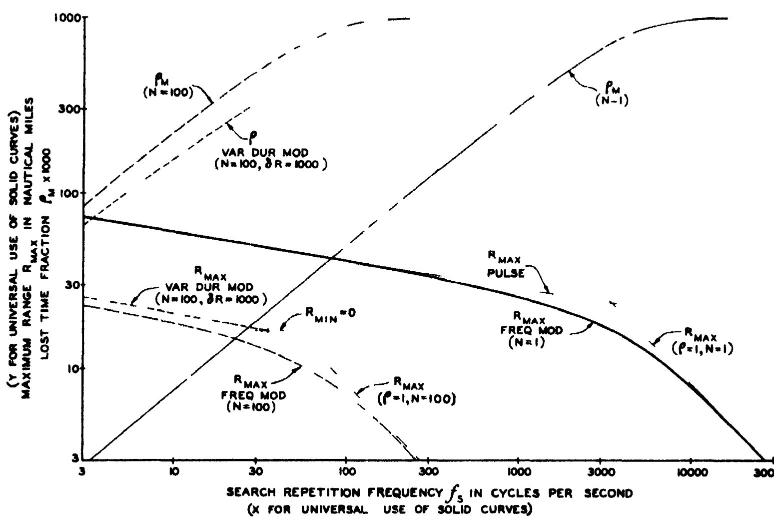


Fig. IX.-1. Noise-limited performance characteristics of various range-only radar search systems, at constant average power.

Maximum range for systems which search all range elements during each single repetition period ($N = 1$) is shown by the full-line curve for R_{max}, and is independent of the number of range elements simultaneously examined. Operation with the number N of pulse repetitions required for a complete search greater than one reduces maximum range. The curve for $N = 100$ is shown dashed in Fig. IX.-1; curves for other values of N would be parallel to those for $N = 1$ and $N = 100$. Comparison of equations (IX.7) and (IX.8) with (IX.4a) and (IX.5a) shows, in fact, that on logarithmic scales the shape of the curve relating R_{max} to f_s does not depend at all upon N, so long as repetition frequency f_r is constant throughout the search; under this condition on f_r, R_{max} is independent of the range resolution δR.

When modulation frequency, or duration of single sweep, varies during the search, maximum range depends not only on search-repetition frequency f_s and number N of elements searched successively, but on width δR of each resolved range element or on minimum range R_{min} as well. Simple generalizations have not been found applicable to operation with variable sweep duration. Maximum range has been determined from equations (IX.21) and (IX.22) for one special case, with $N = 100$ and $\delta R = 1000$ feet, and is shown by the dotted curve for R_{max} in the figure. Since minimum range in a real system must exceed zero, this curve can exist only down to a least maximum range of $(N-\frac{1}{2})\delta R$ if δR remains unchanged throughout the search, as indicated in the figure.

Values of ρ_m, the fraction of the single-sweep duration lost in signal propagation at maximum range, are also plotted in Fig. IX.-1. Each ρ_m curve corresponds to one of the R_{max} curves and is drawn in the same type of line. If data must be gathered with even moderate rapidity by an f-m system which examines a considerable number of range elements in succession, ρ_m will reach values that are not negligible. A penalty is paid in range, and a stiffer penalty is also paid in required width of frequency-modulation band, for this loss of time. In fact, equation (IX.9) shows that as ρ_m approaches unity, the sweep width required for a given range resolution increases very rapidly and soon becomes prohibitive.

Examination of equations (IX.4) and (IX.5), and of similar equations which lead by mere insertion of numerical values to (IX.7) and (IX.8), shows that changes of radio frequency, signal level, noise level, or number of range elements change only the scales used. The form of the relations among f_s, ρ_m, and R_{max} is not affected. In fact, the solid curves and lines of Fig. IX.-1 are universally applicable if their scales are not read as f_s and R_{max} directly, but rather as dimensionless variables x and y, where x signifies $3560(2Nf_1/F_o)^{9/10}\left\{\dfrac{P_t A_s'^2 A_e'}{2\pi F_o \overline{NF}kT}\right\}^{3/10} f_s/f_1$ and y is $22.7(2Nf_1/F_o)^{1/10}\left\{\dfrac{P_t A_s'^2 A_e'}{2\pi F_o \overline{NF}kT}\right\}^{-3/10} R_{max}/\lambda_o$. Any change of the apparatus parameters F_o, P_t, A_s, N, or $\overline{NF}$, or of target area A_e, changes only the coefficients of f_s and R_{max}; on

the logarithmic plot, this acts only to slide the curves along the f_s and R_{max} axes without any change of form.

If ρ_m approaches unity for an average target, such as has been assumed in these comparisons, then signal from a very large target may be discernible above noise at a range sufficient to prevent reflected signal from returning during the same modulation sweep on which it was transmitted. Signals returning during a later sweep of course give erroneous range indications and are to be avoided. Maximum range as limited by maximum permissible time loss rather than noise is $c\rho_m/(2Nf_s)$. This limiting relation is also shown in Fig. IX.-1, for the rather impractical case $\rho_m = 1$, by the steep straight lines so marked. Range limitation by arbitrarily fixed lower values of ρ_m would be represented by straight lines parallel to those for $\rho_m = 1$ but below them.

b. *Other Limitations on Range.* Many other factors than signal/random noise ratio enter into the evaluation of radar performance. Resolution of a weak target adjacent to a strong one, for example, is improved by using a wider receiver noise band than the one giving best ratio of signal to random noise. In the pulse-radar case this permits making fullest use, for target resolution, of the rapid rise and fall of the transmitted pulse. In the f-m case, it permits transient response of the selective channel, shock excited by off-frequency signal from the strong target, to die away before indication of the correctly tuned weak target is initiated on any given frequency sweep; steady-state duration of the desired indication is also increased.

Interfering signals can also limit radar range. An unmodulated continuous-wave signal disturbs the base line of a pulse-radar presentation, in effect enhancing the noise somewhat. It disturbs an f-m system by producing rather sharp beat pulses with the frequency-modulated local mixing signal; these pulses in turn shock excite the selective channels. Discrimination against c-w interference is obtained in the pulse-radar system by the high peak power attainable in the desired pulse. In the f-m case, discrimination is obtained by the fact that the desired signal is present throughout the modulation sweep, while the

interfering signal is only shifted into the selector pass band momentarily, twice per modulation sweep.

With comparable operating speed and resolution, the rejection of c-w interference obtainable appears in principle to be about the same for pulse and for multiple-selector f-m systems (with a possible factor of two in energy favoring the pulse system because of the double-side-band nature of the beat between f-m and c-w signals). In the case of the single-selector f-m systems, discrimination against c-w interference is sharply reduced, presumably to the same degree as signal/noise ratio, by the absence of desired-signal selector excitation during most of the range-search period.

Modulated continuous-wave interfering signals can be much more damaging than unmodulated ones. This is true for any radar system, but the degree of damage depends upon the particular characteristics of the system and of the interfering modulation to such an extent that general conclusions are difficult to draw and are not very significant. It does appear, however, that modulation of the interfering signal by random noise of such character as just to fill the radar r-f channel with strong noise will always be particularly obnoxious.

Pulse interference also acts to limit range. In pulse radar, non-synchronous interfering pulses can be discounted to a remarkable degree by the judgment of a skilled operator. In f-m radar, the narrow band of the selectors acts strongly to reject interference from very short pulses. With a peak-reading indicator, band reduction decreases pulse interference more rapidly than it does random noise. Single-selector f-m systems will be more disturbed by pulse interference than will multiple-channel systems, because of inefficient utilization of the single channel for desired-signal reception. No thorough comparison of multiple-channel f-m radar and pulse radar with respect to pulse interference has been made.

So long as it is prevented from damaging detectors, feed through is important in pulse-radar systems only at minimum range. Strong rejection of low beat frequencies minimizes the disturbance of f-m search systems by feed-through signals, as well as by microphonics, detector

unbalance, and altitude signal. High-frequency modulation of feed-through fields, as by transmitter-tube noise or by loose metal in the vicinity of one antenna, remains damaging and must be carefully avoided.

Sea return is harmful in all systems and must be kept small by concentrating onto the desired target as much as possible of the transmitted energy and by including in the field of view of the receiver as little sea surface as possible. Both these results are attained by using the highest practicable antenna directivity. Multiple-channel f-m systems are less disturbed by sea return than are single-channel ones. Because of their greater range selectivity, even the single-channel search systems see less sea and so are less disturbed by sea return than the broad single-target systems of earlier chapters.

c. *Practical Aspects.* Pulse radar is placed at a practical disadvantage, relatively to frequency-modulated radar, by the necessity of operating transmitters at very high peak-power levels and of preventing damage to the receiver by high feed-through peaks. This is especially true if high average power is to be maintained at low repetition frequencies. Obtaining the very wide r-f bands necessary for resolution of small range differences is technically more difficult with short-pulse than with f-m methods.

Frequency-modulated radar is at a disadvantage because of the complexity of multiple-channel selectors. Single-channel search systems may be acceptably simple, but their ineffective rejection of noise and interference and their slow operation place them at a very serious disadvantage.

Another practical matter is the importance of maintaining a highly uniform rate of change of transmitted frequency during the modulation sweep of an f-m search radar. Very simple considerations, neglecting the effect of transient response of selectors and the time lost in propagation, will serve to illustrate the basic effect of non-linear frequency modulation.

Suppose a radio-frequency band of width W to be swept in modulation during a time interval t_r [equal to $1/(2f_m)$, where f_m is modulation frequency]. Suppose further that during the first half of t_r the transmitted frequency sweeps

uniformly across a band of width $\frac{1}{2}(1+\eta)W$ rather than exactly $\frac{1}{2}W$, covering only the remaining band $\frac{1}{2}(1-\eta)W$ during the second half of the sweep period. Rate of change of frequency $\dot{F}$ will first be constant at $2(1+\eta)Wf_m$, then at $2(1-\eta)Wf_m$. Range frequency for a single target at range R will be $4(1+\eta)Wf_m R/c$ during the first half sweep and $4(1-\eta)Wf_m R/c$ during the second half sweep. Instead of a single range frequency f_0, given by $4Wf_m R/c$, the non-linearity of modulation has produced the two frequencies $f_0+\eta f_0$ and $f_0-\eta f_0$.

For best signal/noise ratio, the noise band Δf of each range-channel selector will be approximately $2f_m$, and beat frequencies differing by $3f_m$ should be indicated as distinct targets separated in range by δR. If the range-frequency difference $8\eta Wf_m R/c$ due to non-linearity becomes $3f_m$, a single target at range R will be shown as two distinct targets on adjacent range channels, with ranges differing by δR. This will occur when the fractional departure η of the sweep from linearity reaches the magnitude $\frac{1}{2}\delta R/R$, which is alternatively expressible as $\frac{3}{4}\Delta f/f_0$. For a practical case, $\delta R/R$ may be $\frac{1}{50}$ and a non-linearity of one per cent will then cause each target to appear definitely on more than one channel. Broadening of target response will begin to appear at about half of this value of η.

Other forms of non-linear sweep, or allowance for filter transients, would modify these numbers, but would not relieve the stringent necessity of maintaining substantially linear sweep if full theoretical resolution is to be attained.

d. *Conclusions.* Single-channel f-m radar search systems are prohibitively slow if high-resolution search in both range and direction is required. Their limited theoretical signal/noise performance is a further disadvantage.

There is little distinction between pulse-radar and multiple-channel f-m radar search systems as to limiting theoretical performance.

At the present stage of practical development, it is definitely simpler to display rapidly the variation of a signal envelope with time than to display rapidly the frequency spectrum of a complex signal. To overcome this

advantage of pulse radar for general search use appears to require much more than the very limited development effort so far expended on f-m search radar.

4. EFFECTS OF SPEED

a. *Direct Effect.* The previous discussion of range and resolution has been simplified by consideration of stationary radar and targets only. The effect of relative motion on f-m search-radar indications must now be examined. This effect is to introduce Doppler frequency shift and so to produce, during modulation upsweep of transmitter frequency, a beat frequency different from that found during modulation downsweep. Range-selector band widths must of course be broadened somewhat to allow for speed effects, but usually not to a sufficient extent to invalidate the conclusions already reached as to order of magnitude of limiting range.

As described in section 4e of Chapter II., moderate Doppler shift decreases the upsweep beat frequency and increases the downsweep frequency when relative motion of radar and target is reducing the range between them. That is, a target moving toward the radar will be indicated at less than actual range during upsweep and at more than actual range during downsweep. On a multiple-channel search radar, each moving target will simply appear twice, at apparent ranges differing by an amount proportional to its speed. Doubling of moving targets can be quite useful in distinguishing them from fixed targets.

In the case of range-scanning search systems using a single selective channel, the speed of search must be halved to avoid confusion by moving targets. That is, one complete modulation cycle rather than a single upward or downward sweep should be devoted to each range channel. Were this not done, moving targets would seem to be single and at the wrong range rather than double and centered at their true ranges.

Targets moving so fast that Doppler shift exceeds frequency displacement due to range will of course give a quite different result. They will also appear doubled, but with the two component indications for each target widely separated. The component separation of each doubled

indication will depend on target range, while the apparent range of the center of the pair will be determined by relative speed of target and radar. This condition should normally be avoided in designing and using radar equipment for range search.

b. *Elimination of Speed of Radar.* Rapid motion of the f-m radar itself, as when airborne, of course results in doubling of all target indications to the same degree. This is usually a decided inconvenience and should be eliminated. Motion of targets themselves would simply alter the component separation caused by motion of the radar.

With range indicated by a cathode-ray beam moving in synchronism with the range scan of the radar, doubling can be eliminated by displacing the cathode-ray beam in range by an amount proportional to speed of radar. For targets ahead of the radar, indicated range would be made to increase during modulation upsweep and to decrease during downsweep. Doubling by motion of targets themselves would remain unaltered, just as if the radar were stationary.

Another way of eliminating target doubling due to radar motion is to alter rate of change of frequency by a suitable amount from upsweep to downsweep. With any of the f-m radar systems, this can be done for targets ahead of the moving radar by slightly decreasing the duration of upsweep and increasing the duration of downsweep. An alteration of average range sensitivity occurs when the effect of radar speed is eliminated by making the modulation cycle unsymmetrical. The slight sensitivity correction required can be made by changing the duration of the complete modulation cycle, or in other ways if preferred.

Let W be the width of modulation sweep, f_R the range-beat frequency for a target at rest relative to the radar and at range R, and f_S a Doppler-beat frequency caused by motion of the radar at speed S. Let t_u be the duration of modulation upsweep and t_d that of downsweep used in the moving radar, and let f_m be the corresponding modulation frequency used in a stationary radar. The conditions

$$f_u = {}^{2WR}/_{(ct_u)} - f_S = f_d = {}^{2WR}/_{(ct_d)} + f_S = f_R = 4Wf_m R/c \qquad \text{(IX.23)}$$

then ensure that range indication shall not be altered by the motion of the radar; f_u and f_d are the total beat beat frequencies for the moving radar during upsweep and downsweep respectively.

If the modulation frequency $1/(t_u+t_d)$ for the moving radar is called f_m', (IX.23) yields

$$f_m' = f_m(1-{f_s^2}/{f_R^2}), \tag{IX.24}$$

$$t_u = \tfrac{1}{2}(1-{f_s}/{f_R})/f_m' = 1/[2f_m(1+{f_s}/{f_R})] , \tag{IX.25}$$

and

$$t_u = \tfrac{1}{2}(1+{f_s}/{f_R})/f_m' = 1/[2f_m(1-{f_s}/{f_R})] . \tag{IX.26}$$

Additive corrections for radar motion are thus independent of sweep width and range, for any system using a single filter tuned to a fixed frequency f_R. Correction *factors* are independent of f_m as well. Only a simple dependence on speed frequency f_s and fixed selector frequency f_R remains. The correction (IX.24) of average modulation frequency from f_m to f_m' is of second order and can usually be disregarded.

c. ***Moving Target Indication*** (*MTI*). Selective indication of moving radar targets with exclusion of confusing fixed targets, known as *MTI*, is a valuable service. *MTI* is accomplished by recording the data obtained during each range search, comparing it in detail with the data obtained on the next range search, and indicating only the difference found. Recordings should be of some temporary sort, lost or erased as soon as used for comparison.

In the case of pulse radar, fully effective *MTI* requires that radio-frequency phase of each transmitted pulse be preserved and compared with the phases of all its target echos. The results of such phase comparison must in turn be preserved and compared for successive pulses. This is accomplished in actual use, but the cost is complicated equipment and critical adjustment.

A very simple way of using f-m radar for *MTI* has been proposed but probably never tried. Since the frequency-modulated transmitter operates continuously, and within each sweep coherently, a radio-frequency phase reference is automatically provided. This is an inherent advantage

of the f-m type of radar for *MTI*. The normal f-m radar beat signal indicates the result of phase comparison between transmitted and target-reflected waves. This comparison is also inherent in the normal f-m radar. The only additional operations required to obtain selective indication of moving targets with f-m radar are the recording of the beat signal for each range search and comparison of it with the beat signal for the following search.

For a single-channel system operating against an isolated target, or for any one channel of a multiple-channel system operating against multiple targets, the beat-note signal has during each sweep substantially a simple sinusoidal form, with growing and decaying portions adjoined thereto by any selective element used. This simple signal is often only of audio or low video frequency, and in the cases mentioned needs merely to be stored for comparison from one modulation cycle to the next and then discarded. This is a relatively easy job of recording, so the total complication to be added to one f-m radar channel for *MTI* is not great.

The discussion of fixed error given in section 2g of Chapter IV. indicates that, with accurately constant range from radar to target, beat-note wave form will repeat accurately for successive modulation cycles. Subtraction from the beat signal currently received of a beat-signal record made during the previous modulation cycle will therefore give exact cancellation for a fixed target, and such targets will not be indicated at all. Target motion through only ¼ of the radio-frequency wave length during each modulation cycle will reverse the beat-note phase for successive cycles, so that subtraction of a record will give very strong indications for such targets. A very sensitive method could thus be provided for distinguishing moving targets from their fixed surroundings.

Great sensitivity to small target motions may also be provided by short-pulse *MTI* radar. This high sensitivity may even prove embarrassing, no target and propagation path being so firmly fixed as to cancel perfectly. The advantage of the f-m *MTI* would lie in the ease of recording the beat signal, for example by a simple magnetic recorder.

5. EXPERIMENTAL SEARCH SYSTEM AN/APQ-19

a. *General Description.* One experimental f-m search radar, designated *AN/APQ-19*, has been built and given limited field tests on the ground. This equipment[5,6] is of a single-channel type with fixed selector tuning, fixed modulation frequency, and range search by variation of modulation sweep (class C_1 of Table IX.-1). Search in azimuth as well as range is provided, with a "B-scope" type of data presentation plotting range against azimuth angle as rectangular coordinates on the screen of an indicator oscilloscope; targets appear as bright points in these coordinates.

Fig. IX.-2 is a block diagram of the *AN/APQ-19* system and Fig. IX.-3 a functional diagram of the circuits used. Supplies for numerous regulated voltages used are conventional and are omitted from the diagrams. Two 18-inch parabolic reflectors from *AN/APS-3* pulse radars are mounted one above the other and rigidly interconnected, so as to scan together in azimuth and tilt together in elevation as a single mechanical unit. A single coaxial-fed dipole antenna is used in each reflector. The transmitting oscillator is an *A-127B* experimental magnetron, frequency modulated internally by helical auxiliary beams of electrons as described in section 4e of Chapter III. The receiver is a simple crystal mixer followed by an amplifier. A conventional intermediate-frequency amplifier is used as a sharply selective beat-note filter, with its rectifier output controlling the spot brightness of a conventional radar-indicator oscilloscope in accordance with the beat-note envelope.

A group of synchronized multivibrators serves as a modulation generator providing several timing outputs, used to control modulation of the transmitter by a special circuit for range search and to control the display of received data. The special circuit for controlling modulation sweep to provide range search is the most novel element of the system, except for the auxiliary-beam magnetron itself, and will be described separately. Provision is made for modifying symmetry of the modulating wave to cancel the effect of motion of the radar itself. Various operating parameters of the system are collected in Table IX.-3.

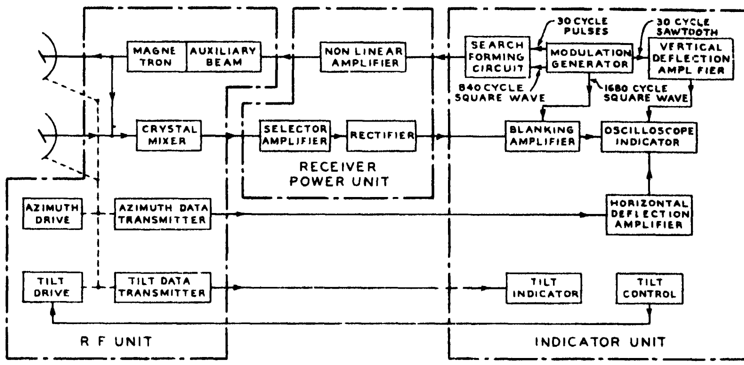


Fig. IX.-2. Block diagram of experimental f-m search radar *AN/APQ-19*.

TABLE IX.-3.

Design and Operating Characteristics of *AN/APQ-19* F-M Search Radar

Transmitted Power	20 watts
Radio Frequency	4000 megacycles per second
Frequency-Modulation Sweep	1 to 5 megacycles per second
Azimuth Sector Searched	150 degrees
Number of Azimuth Elements Searched	30
Azimuth Search Frequency	1 single sweep per second
Time on each Azimuth Element	33 milliseconds
Number of Range Elements Searched	56
Range Search Frequency	30 per second
Modulation Frequency	840 per second
Time on each Range Element	600 microseconds
Time Constant of Selective Channel	140 microseconds
Pass Band of Selective Channel	7.0 kilocycles per second
Center Frequency of Selective Channel	100 or 200 kc./sec.
Minimum Range Searched	1 or 2 miles
Maximum Range Searched	5 or 10 miles
Range Resolution (nominal)	10 or 5 per cent
Speed Sensitivity	13.7 cycles per second per knot

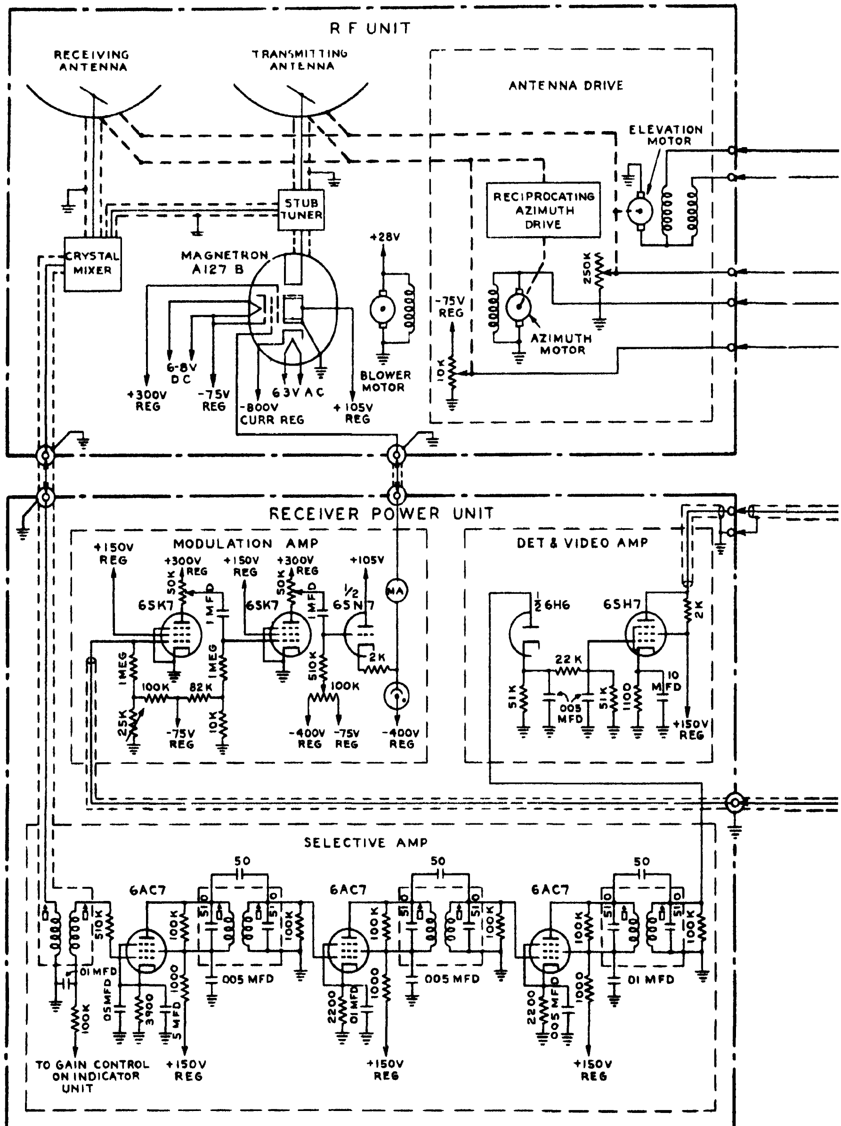
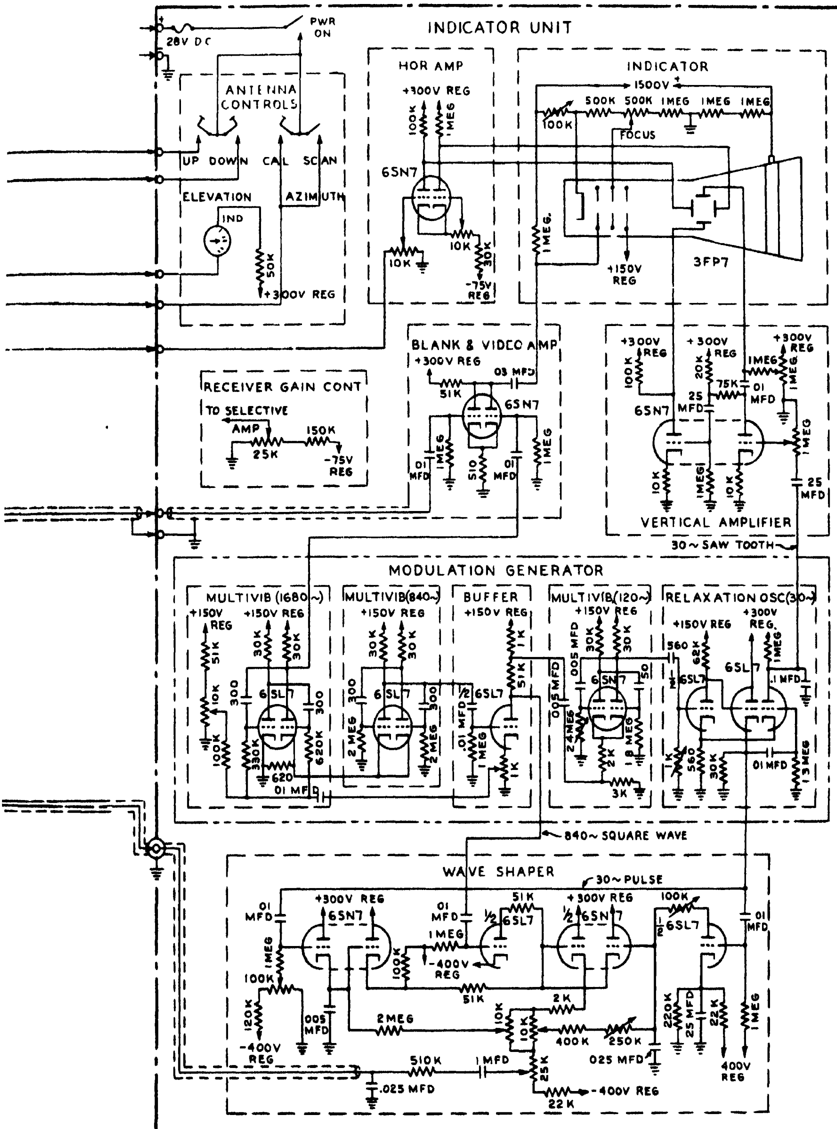


Fig. IX.-3. Functional circuit diagram of

It may be noted that the time lost in propagation at maximum range is $^1/_5$ the total time allotted each range element. The time constant of the selective channel is approximately $^1/_4$ the time per range channel. Allowing twice the filter time constant for build-up and prior-



experimental f-m search radar AN/APQ-19.

element decay transients, over $^1/_4$ of the time per range element remains to provide steady-state filter-output signal for indication. Better operation could in principle be had by short-circuiting the selective channel between successive range elements. Since selection is begun at a

very low signal level to prevent production of ghost targets by inter-modulation, all short-circuiting methods tried have been found to introduce fortuitous transient disturbances greater than the very small unwanted prior-element signals which they suppress.

b. *Production of Modulating Signal.* The timing portion of the modulation and deflection generator used in the *AN/APQ-19* is too conventional to require detailed description. The primary timing element is a free-running symmetrical multivibrator with a frequency of 1680 cycles per second. Actual square-wave modulating signal is derived, through special circuits to be described, from an 840-cycle per second multivibrator synchronized by the 1680-cycle master. Through a buffer amplifier which also supplies the modulation-controlling output, the 840-cycle signal in turn synchronizes a 120-cycle per second multivibrator, which again in its turn synchronizes a final 30-cycle relaxation oscillator. The 30-cycle oscillator produces a sawtooth voltage for rangewise (vertical) deflection of the indicating oscilloscope and, through a cathode follower, a 30-per-second train of short pulses to reset the range-searching circuit of the radar.

There are two items of novelty in this timing system. An 840-cycle square wave of adjustable amplitude is fed back from the buffer amplifier to bias both grids of the 1680-cycle multivibrator. Alternate complete cycles of oscillation of the fast multivibrator are thus made to have somewhat different periods. Alternate halves of the 840-cycle oscillation, corresponding respectively to frequency upsweep and downsweep of the transmitter modulation, may thereby be given different durations as required to compensate for the speed of motion of the radar. Also, 1680-cycle signal is applied to the blanking amplifier which is used to brighten the trace of the indicator in accordance with the strength of the received radar signal. This 1680-cycle signal blanks out the indication during the first half of each single modulation sweep, to prevent showing either the false signal following turn around or the worst portion of the unwanted transient response of the selective amplifier.

Fig. IX.-4 shows the special circuit that is used

to control transmitter modulation so as to provide the stepped-exponential variation of modulation-sweep width [equation (IX.12)] necessary for range search. In the operation of this circuit, each pulse from the 30-cycle tuning oscillator charges C_1 positively and C_2 negatively,

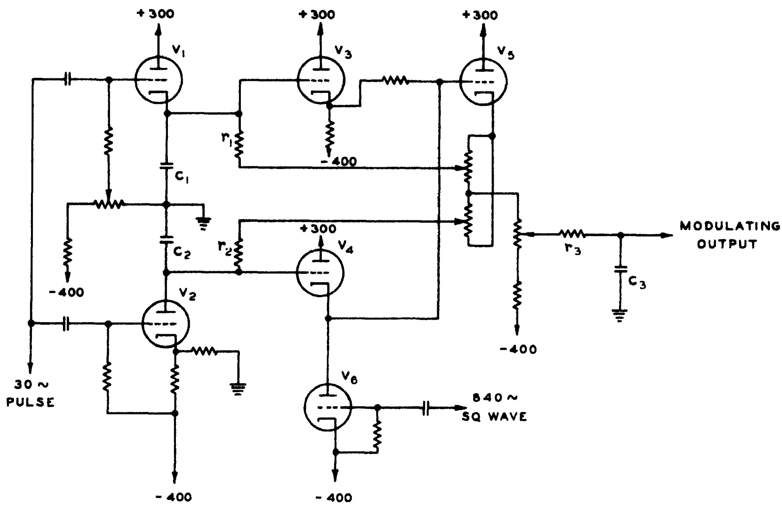


Fig. IX.-4. Circuit for forming modulating signal for range search.

through switching tubes V_1 and V_2 respectively, and so resets the circuit to begin a new range search. Square-wave signal from the 840-cycle multivibrator acts through switching tube V_6 to control the individual frequency-modulation cycles. Tubes V_3, V_4 and V_5 are merely load-isolating cathode followers.

When V_6 is cut off by the 840-cycle square-wave control, cathode follower V_4 is disabled and cathode followers V_3 and V_5 act in cascade to hold the cathode of V_5 at substantially the voltage of C_1. There is then no current through R_1, and C_1 holds its charge. The total voltage of C_1 and C_2 appears across r_2, which discharges C_2 at a corresponding rate. When V_6 is made conducting by the 840-cycle control, V_3 is overpowered because of its high-resistance output connection, so that V_4 and V_5 in cascade hold the cathode of V_5 at substantially the voltage of C_2. There is then no current through r_2, and C_2 holds its charge. The total voltage of C_1 and C_2 appears across r_1,

which discharges C_1 at the corresponding rate. Return points for r_1 and r_2 are adjustably tapped down on the cathode resistor of V_5, as shown, to compensate for bias voltages of V_3, V_4 and V_5. The return points of r_1 and r_2 are thus placed accurately at the corresponding capacitor voltages, so that the capacitor discharge rates are reduced accurately to zero during the time intervals when they should be zero.

Output voltage at the cathode of V_5 is alternately held constant at one level by C_1 during modulation upsweep and at another level by C_2 during downsweep. At all times, whichever capacitor is not in use to hold the output is being discharged, preparatory to setting a new output level on its next cycle of operation. Capacitance, resistance and discharge duration being fixed, each discharge period reduces capacitor voltage by a constant fraction. Output voltage from V_5 then varies, in the interval between search-starting pulses, in just the stepped-exponential way in which it is necessary that modulation-sweep width should vary. Sweep duration is controlled by the 840-cycle multivibrator and sweep width by the operating sequence of C_1 and C_2. Output from V_5 is integrated by r_3, C_3 to provide as final modulating output a sawtooth wave with linear sides but exponentially decaying amplitude. The resulting variation of transmitted frequency with time is shown in Fig. IX.-5.

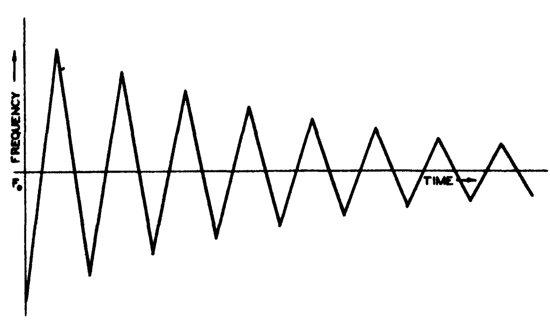


Fig. IX.-5. Frequency-modulation wave form for radar range search.

Conditions to be fulfilled in producing the desired stepped-exponential search follow from the operating sequence of the circuit of Fig. IX.-4. Let the voltage

across C_1 be E' and the voltage across C_2 be E'', both referred to ground. Let capacitor charging be completed with V_6 conducting, so that the first single modulation sweep of each range search takes place upon V_6 being cut off, and let this first sweep be an upsweep of transmitted frequency. Capacitor C_1 then holds its voltage and C_2 discharges during each upsweep; upsweeps are the odd-numbered or $2k+1^{st}$ ones of each range search, where k is any integer [the first sweep ($k=0$) is a special case]. Capacitor C_2 holds and C_1 discharges during each downsweep, and downsweeps are the even-numbered or $2h^{th}$ ones. Voltages at the start of any sweep will be designated by the order number of the sweep as a subscript.

Change in capacitor voltage during any discharging sweep depends on the duration of the sweep, the discharge-circuit time constant, and the difference of the two capacitor voltages at the start of the sweep. Using the notation outlined above, voltage changes are

$$E'_{2k+1} - E'_{2k} = -(E'_{2k} - E''_{2k})\left[1-\epsilon^{-t_d/r_1C_1}\right] = \Delta_k E' \qquad \text{(IX.27)}$$

for C_1 during the $2k^{th}$ sweep and

$$E''_{2h} - E''_{2h-1} = (E'_{2h-1} - E''_{2h-1})\left[1-\epsilon^{-t_u/r_2C_2}\right] = \Delta_h E'' \qquad \text{(IX.28)}$$

for C_2 during the $2h-1^{st}$ sweep, where t_u and t_d are duration of upsweep and downsweep. Net capacitor voltages at any particular time result from the summation of such discharge processes and are given by

$$E'_{2k+1} = E'_{2k+2} = E'_1 + \sum_{j=1}^{j=k} \Delta_j E' \qquad \text{(IX.29)}$$

and

$$E''_{2h} = E''_{2h+1} = E''_1 + \sum_{j=1}^{j=h} \Delta_j \quad , \qquad \text{(IX.30)}$$

where E'_1 and E''_1 are the voltages to which C_1 and C_2 were initially charged at the start of the first sweep.

Substituting from (IX.29) and (IX.30) into (IX.27) and (IX.28), voltage changes on discharge are found to be

$$\Delta_j E' = -(\Delta_j E'' - \Delta_j E')\,u\,(1-d)/(1-ud) \qquad \text{(IX.31)}$$

and

$$\Delta_j E'' = (\Delta_j E'' - \Delta_j E')(1-u)/(1-ud) \ , \qquad \text{(IX.32)}$$

subject to the recursion formula

$$\Delta_j E'' - \Delta_j E' = [(E_1' - E_1'') - \sum_{i=1}^{i=j-1} (\Delta_i E'' - \Delta_i E')](1-ud) \ . \qquad \text{(IX.33)}$$

The time factors $\epsilon^{-t_u/r_2 c_2}$ and $\epsilon^{-t_d/r_1 c_1}$ have here been abbreviated as u and d respectively. By direct application of (IX.28) and (IX.27) to the first and second sweeps,

$$\Delta_1 E'' = (E_1' - E_1'')(1-u) \ , \qquad \text{(IX.34)}$$

$$\Delta_1 E' = -(E_1' - E_1'')(1-d)\, u \ , \qquad \text{(IX.35)}$$

and

$$\Delta_1 E'' - \Delta_1 E' = (E_1' - E_1'')(1-ud) \ . \qquad \text{(IX.36)}$$

Repeated use of (IX.33), starting with the value given by (IX.36), shows that

$$\Delta_j E'' - \Delta_j E' = (E_1' - E_1'')(1-ud)(ud)^{j-1} \ . \qquad \text{(IX.37)}$$

Substitution of this result into (IX.31) and (IX.32), and of $\Delta_j E'$ and $\Delta_j E''$ so found into (IX.29) and (IX.30), gives geometric-series expressions for E'_{2k+1} and E''_{2h}. Summing these series, the expressions

$$E'_{2k+1} = E'_{2k+2} = E_1' - (E_1' - E_1'')u(1-d)(1-u^k d^k)/(1-ud) \qquad \text{(IX.38)}$$

and

$$E''_{2h} = E_{2h+1} = E_1'' + (E_1' - E_1'')(1-u)(1-u^h d^h)/(1-ud) \ . \qquad \text{(IX.39)}$$

are found for the successive voltages held on C_1 and C_2 respectively.

Voltages held during successive upsweeps will have a constant ratio throughout the range search, as is required for proper sweep control, if

$$E_1'/E_1'' = -u(1-d)/(1-u) \ . \qquad \text{(IX.40)}$$

The same condition provides a constant ratio of successive downsweep voltages, and this common ratio is simply ud. Using (IX.40) in (IX.38) and (IX.39),

$$E'_{2k+1} = E'_{2k+2} = E_1'(ud)^k \qquad \text{(IX.41)}$$

and

$$E''_{2h} = E''_{2h+1} = E''_1(ud)^h = -E'_1 u(1-d)(ud)^h/(1-u)\,. \qquad \text{(IX.42)}$$

Referring to equation (IX.12) for the desired variation of sweep width, and noting that successive upsweeps or successive downsweeps are only alternate single sweeps, it is necessary that

$$ud = 1/(1+\delta)^2\,, \qquad \text{(IX.43)}$$

where δ is the desired fractional range change between successive range channels. The voltage swing across integrating capacitor C_3 during the $2k^{th}$ downsweep is $E''_{2k}t_d/(r_3C_3)$, and that during the $2k+1^{st}$ upsweep is $E'_{2k+1}t_u/(r_3C_3)$. These voltage swings are proportional to width of the corresponding modulation sweeps. To maintain the proper ratio $\sqrt{ud}$ between successive single sweeps throughout the range search therefore requires that the additional condition

$$\sqrt{ud}\,E''_1 t_d = -E'_1 t_u \qquad \text{(IX.44)}$$

be met, in view of (IX.41) and (IX.42).

Referring to equations (IX.25) and (IX.26) for the values of t_u and t_d required to cancel a speed frequency f_s in the presence of range frequency f_R, (IX.43) and (IX.44) give

$$E'_1/E''_1 = -\frac{1}{1+\delta}\cdot\frac{f_R+f_s}{f_R-f_s} \qquad \text{(IX.45)}$$

as the required initial ratio of capacitor voltages. With (IX.40) and (IX.43), this provides design expressions for u and d also, and through these still other expressions for $f_m r_1 C_1$ and $f_m r_2 C_2$ are finally found. The exact expressions are rather complicated and not of great general interest. They lead to a somewhat different required value for r_1C_1 from that for r_2C_2, even if upsweep and downsweep have equal duration.

The significant result of this analysis is that the circuit of Fig. IX.-4 is capable of giving just the desired stepped-exponential variation of sweep width shown in Fig. IX.-5, even when speed of the radar itself is to be cancelled. To accomplish this result, with a chosen fractional separation of adjacent range channels, specific values for the two time constants and for the initial

capacitor-voltage ratio can be calculated and must be provided in the circuit. Since upsweep and downsweep give distinct range channels with this circuit, some confusion may result from target motion. To maintain the desired sequence of sweep widths exactly when speed of the radar is cancelled by modifying sweep duration, both time constants and the starting-voltage ratio would also have to be modified. These final speed-cancelling refinements are not included in the *AN/APQ-19*.

Modulation characteristics have been imperfectly linear in the experimental auxiliary-beam magnetrons as used in the f-m search equipment. It is therefore necessary to drive the beam-control grids of the magnetron with a special non-linear amplifier fed from the modulation-forming circuit. Amplifier operating conditions are so adjusted as to secure best overall linearity of frequency modulation. Since the magnetron beam-control grids draw heavy current, they are driven by a cathode follower.

One modulation-forming circuit has been described in detail because it seems typical of the sort of step-acting special circuit likely to be required by any single-filter f-m search radar. Other modulation circuits were also tried, and no doubt a great many others are possible. In particular, photoelectric development of special wave forms by use of a motor-driven optical mask was investigated to some extent. This is a very flexible and stable method of generating special wave forms, such as are required to combat magnetron non-linearity, but is not convenient to adjust for cancellation of a variable radar speed.

c. *Operating Tests*. If azimuth scanning is stopped, maximum theoretical noise-limited range of the *AN/APQ-19* may be estimated by the methods of sections 2 and 3 of this chapter, using data from those sections and from Table IX.-3 of this section. In using equations (IX.7) and (IX.8), or the curves of Fig. IX.-1, allowance must be made from (IX.3) for the transmitter power of the equipment, and from (IX.1) for the fact that the filter used was rather too broad for best signal/noise matching of the beat-note pulse.

Making the above allowances but retaining the receiver

noise figure of 40 and the effective target area of 8000 square meters used in earlier comparisons, maximum noise-limited searchlighting range of the *AN/APQ-19* system should be 31 miles and maximum lost-time fraction 64 per cent. Allowing just the reciprocal band width of the filter for transient decay, good signal for indication should exist for the final $^1/_8$ of each single sweep at maximum range; this means that the indicator grid should be blanked during the first $^7/_8$ of each sweep for clean indications at that range. Ground reflection for air-to-ground operation has been considered in assigning the 8000 square-meter effective area to a medium-sized ship target.

The above results indicate that very good operation should in principle be obtained, even on small targets, out to the full range of 10 miles searched by the experimental equipment, despite the rather unfavorable use made of the filter characteristic at that short range. Azimuth scanning should lead to deterioration of results by an undetermined but probably not very large factor.

Limited field tests were made with a ground installation at the edge of and about 60 feet above a large body of water.[7] A fairly large ship seen broadside was clearly indicated at four miles; smaller ships seen stern-on were lost at ranges of about three miles when receding from the radar. Fixed targets on the far shore, probably large buildings, were clearly and reliably indicated at maximum range of ten miles, which was approximately the radar horizon for the site used. Sea return was only occasionally observed.

Severe microphonics of the magnetron proved to be a limiting factor. This condition resulted from attaching the magnetron too rigidly to the antenna mounting in the experimental apparatus. Azimuth-scanning motion of the antennas therefore excited microphonics in the magnetron, especially at the scan-reversal points. Normal statistical receiver noise never set a limit to operation during the field tests. No airborne tests of this equipment were made.

Non-linearity of magnetron frequency modulation was the most serious limitation encountered. It reduced the effectiveness of desired targets in exciting the filter and greatly reduced the range resolution. Individual fixed

targets at maximum range appeared on two or three successive range channels, while near targets appeared on five or six channels. Increased spreading at short range indicates that greater departure from linearity accompanied wider modulation sweep.

The main conclusion reached from the tests of the experimental *AN/APQ-19* is that much improved linearity of frequency modulation is essential to the development of any successful f-m search radar. Laboratory tests of *A-127B* magnetrons have indicated that operating conditions giving highly linear modulation exist, but such conditions were never attained in the field tests of the complete search system.

The *AN/APQ-19* equipment was also used in a few tests to determine the feasibility of operation with a single antenna for both transmission and reception.[8] With the original two-parabola antenna system, it was found for a very few targets investigated that the strength of received signal depended only to a negligible degree on whether the planes of polarization of transmitting and receiving antenna were parallel or perpendicular. The crossed-polarization single antenna of Fig. III.-5 was then substituted for the two-antenna system and found to produce no significant reduction in signal (other than that resulting from reduced antenna area). These results are far too meager to be conclusive, but are distinctly encouraging with regard to the possibility of eliminating the second antenna from some applications of f-m radar.

6. NOTATION AND REFERENCES

a. *Notation.* The algebraic notation listed alphabetically below has been used in this chapter.

A'_a Effective area of antenna, in square wave lengths, for direction of maximum gain.

A'_e Effective echoing area of target, in square wave lengths, for direction of radar.

c Velocity of propagation of radio waves, 983.23 feet per microsecond.

C_1, C_2, C_3 Capacitors in modulation-forming circuit of the *AN/APQ-19* search system.

d Exponential decay factor of modulation-forming circuit for downsweep.

D Detection factor determining minimum discernible signal.

E'_i, E''_i Voltages across capacitors C_1 and C_2 of modulation-forming circuit at start of i^{th} sweep of range search.

$\Delta_j E'$, $\Delta_j E''$ Changes of voltage across C_1 and C_2 during specific single sweep of modulation.

f_o Fixed frequency to which single selector channel is tuned in certain types of f-m search radar.

f_1 Reference frequency (1670 cycles per second) determining empirical data-integration factor.

f_d Net radar beat frequency developed during modulation downsweep of transmitted radio frequency.

f_m Frequency at which transmitted frequency is modulated.

f'_m Modulation frequency as modified for speed cancellation.

f_r Frequency of repetition of radar pulses, either of radio or of beat frequency.

f_R Radar beat frequency due to range only.

f_s Frequency of repetition of complete range search or scan.

f_S Radar beat frequency due to speed only.

f_u Net radar beat frequency during modulation upsweep of transmitted radio frequency.

Δf Noise band width of single selective channel.

F_o Average radio frequency transmitted.

h, i, j, k Running subscripts identifying individual modulation sweep of range search.

k Boltzmann gas constant, 1.37×10^{-23} joules per degree Centigrade.

K Filter-matching factor determining minimum discernible signal.

M Number of range elements searched during single pulse repetition or modulation sweep.

n Number of data values integrated by observer.

N Number of range elements searched by successive modulation sweeps of complete scan.

$\overline{NF}$ Noise figure of receiver.

P_p R-m-s transmitted power at peak of pulse.

P_r Average received-signal power.

P_t Average transmitted power.

r_1, r_2, r_3 Resistors in modulation-forming circuit of the *AN/APQ-19* search radar.

R	Range or distance between radar and target.
R_{max}	Maximum range searched or observable over noise.
R_{min}	Minimum range searched.
δR	Range difference between targets just resolved, or width of single range element of indicating system.
S	Speed of radar relative to target; also, factor representing statistical criterion for a just-discernible signal.
t_d	Duration of modulation downsweep of transmitted frequency.
t_i	Duration i^{th} modulation sweep of search sequence.
t_p	Duration of radar pulse.
t_r	Duration of single modulation sweep, or $1/f_r$.
t_u	Duration of modulation upsweep of transmitted frequency.
T	Noise temperature of receiving antenna in degrees Centigrade absolute.
T_o	Time interval over which observer is able to integrate repeated data.
T_i	Time during range search at which i^{th} modulation sweep starts.
T_s	Duration of one complete search, or $1/f_s$.
u	Exponential decay factor of modulation-forming circuit for upsweep.
U_{min}	Total energy in pulse just discernible over noise.
U_N	Noise energy, or power per unit frequency band.
U_r	Total energy in received pulse.
$V_1, \ldots V_6$	Vacuum tubes in modulation-forming circuit of the *AN/APQ-19* search radar.
W	Width of band swept in modulation of transmitted frequency.
W_i	Width of band for i^{th} sweep of search sequence.
x	General search-frequency variable.
y	General range variable.
δ	Fractional width of range element searched by radar.
ϵ	Threshold of discernible signal set by threshold of vision; also, base of natural logarithms.
η	Fractional non-linearity of modulation sweep.
λ_o	Wave length of signal at average transmitted frequency.
λ_w	Sweep wave length, or wave length of radio signal having frequency W.
ρ	Fraction of duration of single sweep lost in propagation of signal from radar to target and back.

ρ_m Lost-time fraction at maximum range.

τ Time taken for signal to travel from radar to target and back.

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